roceedings of the I·R·E

JOURNAL of the Theory, Practice, and Applications of Electronics and Electrical Communication

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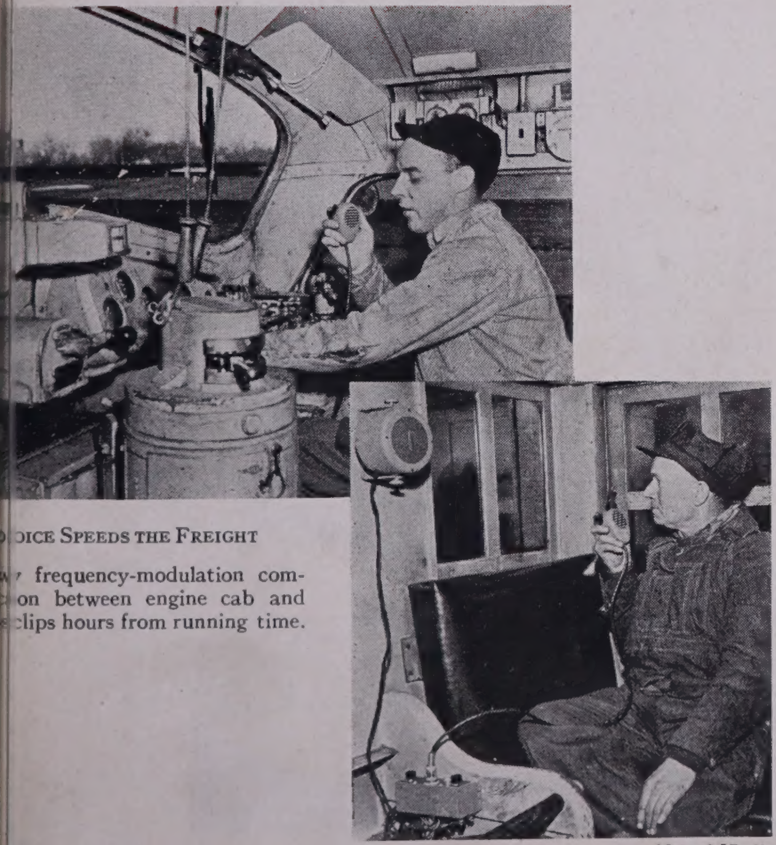

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frequency-modulation com- on between engine cab and clips hours from running time.

Motorola Radio

INTER TECHNICAL MEETING, NEW YORK, N.Y.
January 23, 24, 25, and 26, 1946

DECEMBER, 1945

Volume 33 Number 12, Pt. I

This Issue Published in Two Parts
Part I—Proceedings of the I.R.E.
Part II—Annual Index

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Plastic Materials
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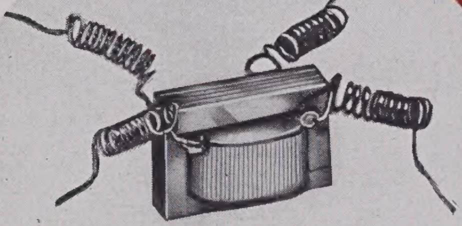

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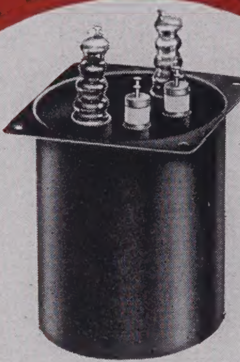

Typical of the special units produced by UTC is this high gain, 100 cycle, matching transformer. Primary impedance 500 ohms, secondary impedance 37,500,000 ohms, shielding suitable for—160 DB signal level.

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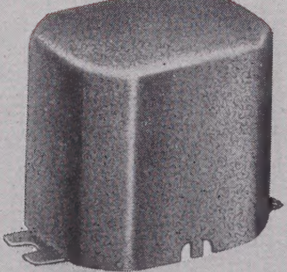
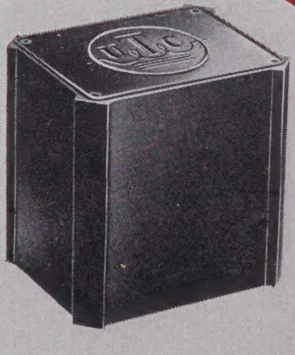


UTC Special Series components cover the entire range of amateur and low priced PA requirements . . . attractively cased . . . economically priced.

FOR THE BROADCAST STATION



UTC linear standard transformers are the ultimate in high fidelity design . . . frequency response guaranteed ± 1.5 DB 20 to 20,000 cycles . . . Low wave form distortion . . . Extremely low hum pickup.

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Standards
on
RADIO WAVE PROPAGATION

DEFINITIONS OF TERMS
RELATING TO GUIDED WAVES

1945

THE INSTITUTE OF RADIO ENGINEERS

The Institute of Radio Engineers

(Incorporated, August 23, 1913)

CONSTITUTION

Adopted at the First Meeting of The Institute of Radio Engineers, May 13, 1912. Amended, November 2, 1914; December 5, 1915; October 7, 1931; March 1, 1939; November 5, 1941; September 8, 1943; and November 29, 1944.

ARTICLE I

NAME AND OBJECT

SEC. 1—The name of this organization shall be The Institute of Radio Engineers, Incorporated.

SEC. 2—Its objects shall be scientific, literary, and educational. Its aims shall include the advancement of the theory and practice of radio, and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members. Among the means to this end shall be the holding of meetings for the reading and discussion of professional papers and the publication of papers, discussions, communications, and such other matters as may be appropriate for the fulfillment of its objects.

ARTICLE II

MEMBERSHIP

SEC. 1—The membership of the Institute shall consist of:

a. Fellows, who shall be entitled to all rights and privileges of the Institute.

b. Senior Members, who shall be entitled to all rights and privileges of the Institute except the right to hold the offices of President and Vice President.

c. Members, who shall be entitled to all rights and privileges of the Institute except the right to hold the offices of President, Vice President, and Director.

d. Associates, who shall be entitled to all rights and privileges of the Institute except the right to hold the offices of President, Vice President, and Director, and the chairmanships of standing Committees and of Sections, and the right to vote. However, Associates of record on March 1, 1939, shall have the right to vote so long as a continuous membership since that date is maintained.

e. Students, who may participate in meetings, wear the badge of the Institute, and receive publications designated by the Board of Directors, but who shall have no other rights and privileges.

SEC. 2—Fellow: For admission or transfer to the grade of Fellow, a candidate shall be at least thirty-two years of age and shall be:

a. An engineer or scientist in radio or allied fields. As such he shall have attained distinction in his profession and shall be eminently qualified to take responsible charge of important radio or allied work. He shall have been in the active practice of his profession for at least ten years, and shall have had responsible charge of important radio or allied work for at least three years, or

b. A professor of engineering or of physical science. As such he shall have attained special distinction as an expounder of the principles of engineering or of science in radio or allied fields. He shall have had at least ten years' experience as a teacher of electrical or physical subjects and shall have had responsible charge, for at least three years, of a course in radio or allied subjects in a school of recognized standing, or

c. A person who has done notable original work contributing to the advancement of engineering or science in radio or allied fields, which has given him a recognized standing at least equivalent to that required for Fellow under paragraph "*a*" or "*b*," or

d. A person regularly engaged in radio or allied work for at least ten years, who, by invention or contribution to the advancement of engineering or science in radio or allied fields, or to the technical literature thereof, has attained a standing at least equivalent to that required for Fellow under paragraph "*a*" or "*b*."

SEC. 3—Senior Member: For admission or transfer to the grade of Senior Member, a candidate shall be at least twenty-six years of age and shall be:

a. An engineer or scientist in radio or allied fields. As such he shall have performed and taken responsibility for important engineering or scientific work in these fields and shall have been in the active practice of his profession for at least six years, or

b. A teacher of radio or allied subjects for at least six years in a school of recognized standing. He shall have been in responsible charge of a major course in such fields, or

c. A person regularly employed in radio or allied work for at least six years, who by invention or by contribution to the advancement of engineering or science in radio or allied fields, or to the technical literature thereof, has attained a standing equivalent to that required for Senior Member under paragraph "*a*," or

d. An executive who, for at least six years, has had under his direction important engineering or research work in radio or allied fields and who is qualified for direct supervision of the technical or scientific features of such activities.

SEC. 4—Member: For admission or transfer to the grade of Member, a candidate shall be at least twenty-four years of age and shall be either:

a. An engineer or scientist in radio or allied fields. As such he shall have demonstrated competence in engineering or scientific work of professional character and shall have been in active practice of his profession for at least three years, or

b. A teacher of radio or allied subjects for at least three years.

SEC. 5—Associate: For admission or transfer to the grade of Associate, a candiate shall be at least eighteen years of age and shall be interested in the theory or practice of radio engineering or of the allied arts and sciences.

SEC. 6—Student: For admission to the grade of Student, a candidate shall be devoting a major portion of his time as a registered student in a regular course of study in engineering or science in a school of recognized standing. Membership in this grade shall not extend more than one and one-half years beyond the termination of his student status as described above.

SEC. 7—The term "allied" as used in this Constitution and Bylaws refers to electrical communication, electronics, and such other technical fields as are directly contributory to or derived from radio.

SEC. 8—The expression "school of recognized standing" includes only schools of college grade providing an engineering or scientific curriculum of not less than four years and granting degrees, and such other schools as may be so designated by the Board of Directors.

SEC. 9—Graduation from a radio or electrical course of a school of recognized standing shall be accepted as equivalent to one year's experience in radio or allied fields. Full-time graduate work, or part-time graduate work with teaching, in radio or allied courses in a school of recognized standing, shall be accepted as equivalent to professional experience, for a period not exceeding two years.

SEC. 10—The time requirements for admission to any grade of membership may be satisfied by applying *pro rata* the experience of the candidate under the various alternative requirements for that grade.

SEC. 11—The terms "member" and "membership" when printed without an initial capital where used in this Constitution and Bylaws include all grades.

SEC. 12—The term "voting member" where used in this Constitution and Bylaws means a member entitled to vote on Institute matters.

ARTICLE III

ADMISSIONS, TRANSFERS, AND EXPULSIONS

SEC. 1—Admission or transfer to Fellow grade shall be by invitation by the Board of Directors only.

SEC. 2—Applications for admission or transfer to any grade of membership, except Fellow, shall be submitted to the Board of Directors. An affirmative vote of at least two thirds of the Board members voting shall elect or transfer an applicant to any grade.

SEC. 3—A reapplication for admission or transfer may be made after the expiration of one year from the date of a rejection.

SEC. 4—The admission fee and dues are payable on notification of election and if not received within six months from notification, the election shall be considered void.

SEC. 5—A member in good standing may resign by submitting a written resignation to the Secretary.

SEC. 6—Subject to the approval of the Board of Directors, a resigned member may resume his membership upon payment of current dues.

SEC. 7—When a member's dues become four months in arrears his membership shall terminate. Subject to the approval of the Board of

Directors, membership may be resumed on payment of all dues in arrears or on payment of a new entrance fee and current dues.

For purposes of Article II, Section 1-d, resumption of membership as provided under Sections 6 and 7 of this Article shall not constitute maintenance of continuous membership.

Sec. 8—To initiate action toward expulsion of a member, a written complaint must be submitted to the Board of Directors, which, if it deems the reason sufficient, shall notify the accused by letter of the charges against him and of the place and date of the hearing, which shall be at least twenty days away. The accused may present his defense in person, in writing, or by an authorized representative. There shall be a majority of the members of the Board of Directors present at the hearing and the votes cast must be unanimous in order to expel. The action of the Board of Directors shall be final and conclusive.

ARTICLE IV

ENTRANCE FEES AND DUES

Sec. 1—The entrance fees, transfer fees, and annual dues shall be as follows:

Entrance Fees	
Fellow	$ 5.00
Senior Member	5.00
Member	3.00
Associate	3.00
Student	None

The transfer fee from one grade of membership to another shall be the difference between the corresponding entrance fees except that there shall be no fee when transferring immediately from Student to Associate membership.

Annual Dues	
Fellow	$10.00
Senior Member	10.00
Member	6.00
Associate	6.00
Student	3.00

Sec. 2—The annual dues shall be payable in advance on the first day of January.

Sec. 3—Under exceptional circumstances, the payment of fees and dues may be deferred or waived in whole or in part by the Board of Directors.

ARTICLE V

OFFICERS

Sec. 1—The governing body of the Institute shall be the Board of Directors and shall consist of the President, Vice President, Secretary, Treasurer, Editor, nine elected Directors, five appointed Directors, and the two most recent past Presidents.

Sec. 2—Except for the elected Directors, the terms of all officers shall be for one year each.

Sec. 3—The terms of the elected Directors shall be for three years each.

Sec. 4—Each year of a term of office established in Article V shall begin with the assembly of the Board of Directors at its annual meeting and terminate with the assembly of the Board of Directors at its following annual meeting.

Sec. 5—No officer shall receive, directly or indirectly, any salary, compensation, or emolument from the Institute, either as such officer, or in any other capacity, unless authorized by a vote of a majority of the entire Board of Directors, except as authorized by the Bylaws.

ARTICLE VI

MANAGEMENT

Sec. 1—The President shall be the regular presiding officer at meetings of the Board of Directors and at meetings of the Institute. He shall be an *ex officio* member of each committee.

The Vice President shall assume the duties of the President in the absence or incapacity of the President.

In the event that neither the President nor the Vice President can personally act, the Board of Directors may elect a chairman from its membership who is authorized to perform the presidential duties during the period of the incapacity of the President and Vice President. The tenure of such temporary chairman shall be at the discretion of the Board of Directors.

Sec. 2—The Board of Directors shall manage the affairs of the Institute. An annual report shall be made to the members on the activities and finances of the Institute.

Six members of the Board of Directors shall constitute a quorum.

Sec. 3—The Board of Directors may make, amend, or revoke Bylaws to this Constitution. The proposed changes and reasons therefore shall be mailed to all members of the Board at least twenty days before the stipulated meeting at which the vote shall be taken. Two thirds of all votes received at the stipulated meeting shall be required to approve any new Bylaw, amendment, or revocation.

Sec. 4—The Secretary shall attend all meetings of the Board of Directors and principal meetings of the Institute and prepare the business and record the proceedings thereof. He shall have charge of the books of account of the Institute, and shall furnish from them such information as is requested by the Board of Directors. He shall conduct the correspondence of the Institute and keep full records thereof.

The Institute shall secure a surety bond on the Secretary. An annual audit of the affairs of the Institute shall be made by certified public accountants and submitted to the Board.

Sec. 5—The Treasurer, under the control of the Board of Directors, shall have general supervision of the fiscal affairs of the Institute.

Sec. 6—All funds received by the Institute shall be deposited in an account requiring the signatures of at least two of the following for withdrawal: President, Vice President, Secretary, Treasurer, and Editor. Funds from this account may be deposited in other accounts authorized and limited in size by the Board of Directors. Funds from these accounts shall be withdrawable on the signatures of authorized bonded individuals for current disbursements. Before additional funds are transferred from the first-mentioned account to another account, the individuals responsible for such other accounts shall submit a statement of the disposition of the previously expended funds to the Treasurer.

Sec. 7—All committees shall be appointed by the Board of Directors.

Sec. 8—The fiscal year of the Institute shall end with the thirty-first day of December.

ARTICLE VII

NOMINATION AND ELECTION OF PRESIDENT, VICE PRESIDENT, AND THREE DIRECTORS, AND APPOINTMENT OF SECRETARY, TREASURER, EDITOR, AND FIVE DIRECTORS

Sec. 1—On or before July first of each year, the Board of Directors shall submit to qualified voters a list of nominations containing at least one name each for the office of President and Vice President and at least six names for the office of elected Director and shall call for nominations by petition.

Nominations by petition may be made by letter to the Board of Directors setting forth the name of the proposed candidate and the office for which it is desired he be nominated. For acceptance a letter of petition must reach the executive office before twelve o'clock noon on the last week-day prior to August fifteenth of any year and shall be signed by at least thirty-five voting members.

Each proposed nominee shall be consulted and if he so requests his name shall be withdrawn. The names of proposed nominees who are not eligible under the Constitution shall be withdrawn by the Board.

On or before September first, the Board of Directors shall submit to the voting members as of August fifteenth, a list of nominees for the offices of President, Vice President, and elected Director, the names of the nominees for each office being arranged in alphabetical order. The ballots shall carry a statement to the effect that the order of the names is alphabetical for convenience only and indicates no preference.

Voting members shall vote for the candidates whose names appear on the list of nominees, by written ballots in plain sealed envelopes, enclosed within mailing envelopes marked "Ballot" and bearing the member's written signature. No ballots within unsigned outer envelopes shall be counted. No votes by proxy shall be counted. Only ballots arriving at the executive office before twelve o'clock noon on the last week-day prior to October twenty-fifth shall be counted. Ballots shall be checked, opened, and counted under the supervision of the Tellers Committee between October twenty-fifth and the first Wednesday in November. The result of the count shall be reported to the Board of Directors at its first meeting in November and the nominees for President and Vice President and the three nominees for Director receiving the greatest number of votes shall be declared elected. In the event of a tie vote the Board shall choose between the nominees involved.

Sec. 2—The five appointed Directors, Secretary, Treasurer, and Editor shall be appointed by the Board of Directors at its annual meeting to serve until the next annual meeting.

Sec. 3—The Board of Directors is authorized to fill a vacancy occurring in the governing body.

ARTICLE VIII

MEETINGS

Sec. 1—There shall be an annual meeting of the Board of Directors during January of each year at which newly elected officers shall begin their terms of service, and the Board shall make necessary appointments. There shall be a meeting of the Board of Directors in November on or after the first Wednesday to receive the report of the Tellers Committee.

Sec. 2—There shall be an annual meeting of the Institute as soon as practicable after the annual meeting of the Board of Directors at which general reports of the Secretary and Treasurer shall be presented.

Sec. 3—Meetings of the Board may be held at such times as are necessary to carry out the provision of this Constitution and shall be held at such other times as any five members of the Board may determine, but only on notice to all members of the Board.

ARTICLE IX

SECTIONS AND OTHER GROUPS

Sec. 1—The Board of Directors may authorize the establishment of sections and other groups of members for the purpose of promoting the interests of the Institute. The Board of Directors may, at its discretion, terminate the existence of any such group.

ARTICLE X

AMENDMENTS

Sec. 1—Amendments to this Constitution may be proposed by means of a resolution adopted by the Board of Directors or by means of a petition signed by at least thirty-five voting members. Such proposed amendment or amendments shall be submitted to legal counsel by the Board of Directors, and, if in the opinion of such counsel, they are in accordance with the laws under which the Institute is organized a copy shall be mailed with a letter ballot to each member.

Sec. 2—Constitutional amendment ballots shall be mailed to the voting members at least sixty days before the date appointed for counting the ballots and the ballots shall carry a statement of the time limit for their return to the executive office. The ballots after marking shall be placed in plain sealed envelopes, enclosed within mailing envelopes marked "Ballot" and bearing the member's written signature. Only ballots within signed outer envelopes shall be counted. No votes by proxy shall be counted. Only ballots arriving at the executive office prior to the stated time limit shall be counted. The Tellers Committee shall count such votes and report to the Board of Directors at its next meeting. If the total vote be at least twenty per cent of the total voting membership and if at least seventy-five per cent of all votes cast shall be favorable, the proposed amendment or amendments shall become part of this Constitution.

Sec. 3—Amendments shall take effect thirty days after their adoption, but officers and officers-elect of the Institute at the time any amendment becomes effective shall continue in office until the end of the terms for which they were elected.

Sec. 4—Copies of the amendments shall be distributed to the members as soon as practicable after adoption.

Sec. 5—A complete history of amendments shall be kept in the files of the Institute.

Bylaws

(As of May 5, 1945)

Article VI, Section 3, of the Institute Constitution provides for Bylaws as follows:

"The Board of Directors may make, amend, or revoke Bylaws to this Constitution. The proposed changes and reasons therefor shall be mailed to all members of the Board at least twenty days before the stipulated meeting at which the vote shall be taken. Two thirds of all votes received by the stipulated meeting shall be required to approve any new Bylaw, amendment, or revocation."

MEMBERSHIP

Sec. 1—Institute members are authorized to use the following abbreviations or symbols indicating their grade of membership:

Fellow—F.I.R.E.
Senior Member—S.M.I.R.E.
Member—M.I.R.E.
Associate—A.I.R.E.

Sec. 2—The emblem of the Institute is copyrighted and shall be reproduced only in connection with official business of the Institute.

Sec. 3—Applicants for membership shall furnish names of references as follows:

For Senior Member—Five Fellows or Senior Members.
For Member—Four Fellows, Senior Members, or Members.
For Associate—Three Fellows, Senior Members, Members, Associates, or other responsible individuals.
For Student—A member of the faculty of his school.

Sec. 4—When the work or location of an applicant for Senior Member or Member grade is such as to make impracticable compliance with Section 3, the Admissions Committee may waive that section upon obtaining other suitable references.

Sec. 5—The name of each newly admitted or transferred Senior Member, Member, or Associate shall be published in the Proceedings.

Sec. 6—A notice of his election shall be sent to each newly admitted member together with a bill for his entrance fee and dues, if not previously paid, dues being computed for the remainder of the calendar year beginning with the quarter next succeeding the date of his election. His entrance fee or dues remaining unpaid, additional bills shall be sent the newly elected member sixty days and one hundred and twenty days after notification of his election, in the last instance accompanied by a warning that his election will be considered void if his admission fee and dues are not received within six months of his notification of election.

The names of all individuals elected to membership who fail to pay admission fees or dues within six months of notification of election shall be turned over to the Membership Committee.

Sec. 7—The name of each applicant for admission to the Institute except in the case of Senior Member or Member grade, shall be placed on the mailing list for the Proceedings immediately upon admission to membership and the receipt of his entrance fee and dues.

Sec. 8—The name of each applicant for admission to the grade of Senior Member or Member shall be placed on the mailing list for the Proceedings immediately upon receipt of his application and the entrance fee.

Sec. 9—Proceedings will be sent to members for a period not shorter than that covered by the payment of dues. Upon request, a newly admitted member may receive the Proceedings for previous quarters of the current year provided dues are remitted for such additional period.

Sec. 10—Admission or transfer to any grade except Fellow may be proposed by any member acting as sponsor, or by the Membership Committee, by supplying to the Admissions Committee sufficient information and testimonials from the required number of references to satisfy the Admissions Committee as to qualifications. Such proposals shall be acted upon by the Admissions Committee, and, if approved, transmitted to the Board of Directors for their action. If approved by the Board of Directors, an Invitation Blank shall be sent to the proposed member inviting him to accept the grade of membership proposed which membership grade shall become effective automatically and immediately upon his supplying the biographical and professional information required, and paying the necessary dues and fees. The name of an invitee shall be placed on the mailing list for the Proceedings immediately upon receipt of dues and fees. The proposal and invitation blanks shall be drawn up by the Membership Committee so as to avoid unconstitutional action due to the sponsor supplying incorrect information.

Sec. 11—The Membership Committee may recommend for transfer to higher grade those members who they think are qualified.

Sec. 12—Each year, the Awards Committee shall recommend to the Board of Directors nominees for Fellow grade. A citation summarizing the accomplishments of the nominee shall be a part of each recommendation.

Sec. 13—Diplomas shall be presented to the newly elected Fellows. If practicable, this presentation shall be made by the President at the next Annual Convention.

Sec. 14—A bill shall be sent to each member not later than December 31 covering his dues for the following year. A second bill shall be mailed on or about March first to each member whose dues remain unpaid. Not later than April first, each member whose dues remain unpaid shall be so notified by the Secretary and informed that, in accordance with Article III, Section 7, of the Constitution, should his dues remain unpaid after April 30, his membership will terminate and he will lose the right to vote and to receive the publications of the Institute. On May first, the name of each member whose dues remain unpaid shall be removed from the roll of membership and such member shall be sent a notice to the effect that according to Article III, Section 7, his membership in the Institute has in fact terminated. The list of such terminated memberships shall be turned over to the Membership Committee.

Sec. 15—The mailing of bills or statements to the last known address of a member shall be considered a valid notice of indebtedness.

Sec. 16—On resuming membership and paying dues in arrears, a member may receive available copies of the Proceedings during the period covered by the back dues. A rebate of 25 cents per copy will be made in lieu of copies of the Proceedings not available.

Sec. 17—The Board, by this Bylaw, waives the dues of each and every member of the Institute who has attained the age of 65 years and has been a member of the Institute for 35 years.

SEC. 18—The Secretary shall, as soon as possible after the end of each year, present to the Board of Directors a report giving the number of members of each grade at the close of the year, and the gains and losses in each grade of membership during the year as a result of elections, transfers, reinstatements, debts, resignations, failure to pay dues, and all other causes. This report shall also give the number of newly elected members whose elections become void during the year through the failure to pay entrance fee and initial dues, and the number of newly elected members who at the end of the year have not yet paid entrance fee and initial dues but whose election is still effective as of that date.

BOARD OF DIRECTORS

SEC. 19—Regular meetings of the Board of Directors shall be held in the office of the Institute in New York, N. Y., on the first Wednesday of each month except in July and August, unless otherwise set as provided for in Section 20.

SEC. 20—The time or place of a regular Board meeting may be altered or the meeting canceled only by majority vote of all Board members not less than twenty days before the original date or the new date set for the meeting, whichever is the earlier. Notice of such change shall be given all Board members by mailing notice not less than ten days before the original or the new date, whichever is the earlier.

SEC. 21—As provided in the Constitution, special meetings of the Board of Directors may be called by any five members of the Board on notice to all other members. Notice of such special meeting, giving the time and place of meeting, the purpose of the meeting, and the names of the Board members calling the meeting shall be mailed to all Board members not less than twenty days before the date set for the special meeting. The place of such special meetings may be only the Institute office in New York, N. Y., unless otherwise authorized by a majority vote of all Board members.

SEC. 22—The Secretary, appointed as prescribed in Article VII, Section 2, of the Constitution, shall be paid a salary determined by the Board of Directors.

EXECUTIVE COMMITTEE

SEC. 23—The Board of Directors in managing the affairs of the Institute according to the Constitution shall appoint a committee of Officers and members of the Board of Directors called the Executive Committee to exercise powers and assume duties of executive management to the extent that the Board of Directors directs and as outlined in Bylaws subject to the following limitations: that the Board may by a majority vote of those present at any meeting overrule any act or decision of the Executive Committee, except insofar as any act has in fact been carried out, or suspend at any meeting any power conferred upon the Executive Committee, such suspension to remain in effect pending repeal of the Bylaw conferring such power, or direct any action or plan of the Executive Committee.

SEC. 24—The Executive Committee shall direct and co-ordinate the work of all standing committees except Appointments, Awards, Constitution and Laws, Executive, Investments, Nominations, and Tellers, unless the Board of Directors directs otherwise.

SEC. 25—The Executive Committee shall approve all membership applications below Fellow grade if and when such applications comply with entrance or transfer routines specified in the Constitution and Bylaws. The Executive Committee shall handle readmissions on the basis of policies established by the Board.

SEC. 26—The Executive Committee shall direct the formulation of and present the annual budget to the Board of Directors for approval.

SEC. 27—The Executive Committee shall manage the office of the Institute, which will include publications, advertising, and the setting of salaries (with the exception of Officers' salaries), within the limits of the budget.

SEC. 28—The Executive Committee shall set dates and locations of conventions, conferences, and meetings of the Institute and shall manage them.

SEC. 29—The Executive Committee shall supervise the formation and operation of Sections and propose their approval to the Board of Directors.

SEC. 30—The Executive Committee shall direct and manage Institute standardization activities including relations with other organizations.

SEC. 31—The Executive Committee may divide its duties and responsibilities among its members as it sees fit and conduct its business as it finds necessary, and shall formulate a "Code of Administrative Practice."

SEC. 32—Minutes of the meetings of the Executive Committee shall be sent to the members of the Board of Directors.

SEC. 33—In addition to its other duties, the Executive Committee is charged with broadly considering Institute policies, and making appropriate recommendations to the Board of Directors on its own initiative.

NOMINATIONS

SEC. 34—The Secretary shall circularize all Section chairmen and secretaries before March first of each year requesting the submission of suitable names to be considered by the Nominations Committee for the various elective offices.

SEC. 35—The Nominating Committee shall submit more than one name for each elective office.

SEC. 36—The Nominating Committee shall transmit its list of proposed nominees to members of the Board of Directors at least a week before the date at which the Board is expected to act upon them.

SEC. 37—When the Board of Directors has not scheduled a meeting to be held between August 14 and September 1 of each year, the Executive Committee is authorized to take any necessary actions on any petitions and to approve the ballot.

SECTIONS

SEC. 38—A petition for the formation of a Section shall be signed by not fewer than twenty-five (25) Fellows, Senior Members, Members, and Associates with mailing addresses within the proposed territorial limits.

SEC. 39—The territory of a Section shall be specified by the Board of Directors.

SEC. 40—All Sections shall accept and conform to a "Constitution for Sections" provided by the Institute Board of Directors.

SEC. 41—For Section maintenance, the Institute shall pay to each Section for each calendar year the following sums:

Seventy-five cents per member for each member up to and including 700 and ninety cents per member for each member over 700 plus ten dollars per meeting for not more than ten meetings.

"Member" means Fellows, Senior Members, Members, and Associates with mailing addresses within the territory of the Section, on December 31st of the calendar year for which payment is made, and "meeting" means meetings of the Section within the calendar year.

SEC. 42—Sections shall have no authority to contract debts for, pledge the credit of, or in any way bind the Institute.

SEC. 43—Section Secretaries shall forward to the Secretary of the Institute a report of each meeting held by the Section for the presentation or discussion of papers, and during January of each year a financial statement for the preceding year.

SEC. 44—A Section of the Institute may co-operate with other organizations in the holding of joint meetings and may invite members of such organizations and the public to its meetings.

SEC. 45—Failure of a Section to maintain the required activities, which shall include the holding of at least five meetings each year and also the maintenance of a minimum membership of 25 Fellows, Senior Members, Members, and Associates, and shall place the Section on probation. All members of the Section shall be informed of the probation by the Secretary of the Institute who shall also call to their attention the requirements for maintaining the Section.

If the delinquency continues for a second year, a second notification to the Section membership shall be made by the Institute Secretary and the Board of Directors shall be informed of the probationary status of the Section.

If the delinquency continues for a third year, the Section shall, thereupon, be dissolved. The Secretary shall so report to the Board of Directors and so inform the Section membership.

The Board of Directors is authorized to waive, in whole or in part, the application in any particular case of the contents of this Bylaw during war and six months thereafter.

COMMITTEES AND REPRESENTATIVES

SEC. 46—The standing committees, each of which shall normally consist of five or more persons, shall include the following:

Admissions
Appointments
Awards
Board of Editors
Constitution and Laws
Education
Executive Committee of the
  Board of Directors
Investments
Membership
Nominations
Papers
Public Relations
Sections
Tellers

Annual Review
Antennas
Circuits
Electroacoustics
Electronics
Facsimile
Frequency Modulation
Radio Receivers
Radio Transmitters
Radio Wave Propagation
Standards
Symbols
Television

These committees shall be advisory to the Board of Directors on those matters which are reasonably described by the committee names, except as defined in these Bylaws.

The terms of appointments of the Admissions, Awards, Board of

Editors, Constitution and Laws, Education, Investment, Membership, Nominations, Papers, Public Relations, Sections, and Tellers Committees shall start with the first day of the month following appointment and continue until the date the succeeding terms of appointments take effect. The Board may specifically advance or delay the terminating date of any committee and the starting date of a succeeding committee. The Board shall make appointments to the following committees: Annual Review, Antennas, Circuits, Electroacoustics, Electronics, Facsimile, Frequency Modulation, Radio Receivers, Radio Transmitters, Radio Wave Propagation, Standards, Symbols, and Television, each year between January first and May first and the terms of appointments shall be from May first of the year when the appointments are made until April thirtieth of the following year. Additional appointments may be made to fill vacancies or to care for special cases as the need arises, with the term of the appointment expiring April thirtieth.

Sec. 47—The Membership Committee shall include the Chairman of the Membership Committee of each Section, *ex-officio*.

Sec. 48—The Sections Committee shall include the Chairman of each Section, *ex-officio*.

Sec. 49—The Appointments Committee shall be appointed by the Board of Directors at its regular November meeting. Its membership shall be chosen from the continuing and newly elected members of the Board of Directors. At least fourteen days before the next annual meeting of the Board of Directors, the Appointments Committee shall mail to the members of the new Board of Directors a list of candidates for the offices of Secretary, Treasurer, Editor, and Appointed Directors, and Chairmen and members of the Awards, Constitution and Laws, Executive, Nominations, and Tellers Committees.

Sec. 50—The Executive Committee shall appoint all standing committees except the Appointments, Awards, Constitution and Laws, Executive, Investments, Nominations and Tellers Committees.

Sec. 51—The President, the Secretary, the Treasurer, the Editor, and at least two other members of the Board of Directors shall comprise the Executive Committee. The President shall be Chairman, the Treasurer shall be Vice Chairman, and the Secretary shall be Secretary of the Executive Committee. The terms of appointments of all members of the Executive Committee, except the President, shall start with the first meeting of the Committee after appointment and shall continue notwithstanding any change in status on the Board until the first meeting of the succeeding Executive Committee.

Sec. 52—The Investments Committee shall consist of the Treasurer as Chairman, the President, the Secretary, with such additional members as the Board may wish to appoint.

The Investments Committee is authorized to buy and sell securities in the name of the Institute by means of an order upon the Custodian of Institute Securities, signed by two members of the Investments Committee authorized to do so by the Board of Directors, directing the Custodian to sell securities owned by the Institute and to deposit the proceeds in the Investments Account, or ordering the Custodian to buy securities and to charge the cost to the Investments Account, such account to be maintained in the Custodian's institution. Interest and dividends received from securities shall be deposited in the Investments Account. The moneys in the Investments Account is to be subject to withdrawal by the Board through its regularly constituted officers. The identity of special funds shall be maintained when changes in securities are made. A report on sales and purchases of securities shall be made to the Board by the Investments Committee at the Board's first meeting after each transaction is consummated.

Meetings of the Investments Committee may be called on two days notice by mail, or on one-day notice by telephone or telegraph, by the Chairman, or in his absence or disability, by any member of the committee. A quorum shall consist of two members of the committee.

Sec. 53—The President, with the consent of the Board of Directors is authorized to appoint teachers of science or engineering who are Institute members as Institute Representatives. Each such Institute Representative is charged with promoting the welfare of the Institute at his school, particularly in matters relating to Student membership.

Sec. 54—The Board of Directors may appoint representatives of the Institute on joint committees, boards, and other local, national, and international bodies.

PUBLICATIONS

Sec. 55—The Institute shall publish a periodical under the title, Proceedings of the Institute of Radio Engineers. The Board of Directors may establish additional publications either periodical or otherwise in character.

Sec. 56—The Secretary is authorized to receive annual subscriptions to the monthly Proceedings at the rate of ten ($10.00) dollars per annum with an extra postage charge when the bulk rate of postage does not apply. A discount of fifty per cent from the subscription price of ten ($10.00) dollars will be allowed to colleges and public libraries upon direct subscription to Institute headquarters. Members, publishers, and subscription agencies may be allowed a discount of twenty-five per cent.

Sec. 57—The price of a single annual subscription to the Proceedings for a Fellow, Senior Member, Member, or Associate shall be $5.00, to be included in his annual dues as specified in Article IV of the Constitution.

Sec. 58—The price of copies of the Proceedings supplied to a newly admitted Fellow, Senior Member, Member, or Associate in advance of the period for which dues are payable shall be included in his entrance fee which is specified in Article IV of the Constitution.

INDEX TO CONSTITUTION AND BYLAWS

Roman numerals refer to Articles; arabic numerals, to Sections; and numbers in parentheses, to Bylaws.

- Abbreviations.....(1)
- Administrative Practice Code.....(31)
- Admission.....III, 1, 2, (10)
  - Associate.....II, 5
    - Posting.....(5)
  - Member.....II, 4
    - Posting.....(5)
  - Senior Member.....II, 3
    - Posting.....(5)
  - Student.....II, 6
  - Time Requirements.....II, 10
- Admissions Committee.....(46)
- Advertising.....(27)
- "Allied".....II, 7
- Amendments
  - Bylaws.....VI, 3
  - Constitution.....X, 1
    - Ballot.....X, 2
    - Distribution.....X, 4
    - Effective Date.....X, 3
    - File.....X, 5
- Annual Review Committee.....(46)
- Antennas Committee.....(46)
- Applications.....III, 1, 2; (25)
  - Reapplication.....III, 3
- Appointed Directors.....V, 1; (49)
- Appointments Committee.....(46), (49), (50)
- Associate
  - Abbreviation.....(1)
- Associate (continued)
  - Dues.....IV, 1
  - Entrance Fee.....IV, 1
  - Posting of Names.....(5)
  - Privileges.....II, 1
  - Proceedings
    - Back Copies.....(58)
    - Subscription Rate.....(57)
  - Reference.....(3)
  - Requirements.....II, 5
  - Symbol.....(1)
  - Transfer
    - Fee.....IV, 1
    - To Higher Grade.....(10)
- Awards Committee.....(46)
  - Appointment.....(49)
  - Fellow Nominees.....(12)
- Bills.....(15)
  - Dues.....(14)
  - New Members.....(6)
- Board of Directors.....V, 1
  - Cancellation of Meeting.....(20)
  - Chairman.....VI, 1
  - Change in Meeting Place.....(20)
  - Change in Meeting Time.....(20)
  - Duties.....VI, 2
  - Meetings.....VIII, 1, 3; (19)
    - Annual.....VIII, 1
    - Quorum.....VI, 2
    - Regular Date.....(19)

- Board of Directors (continued)
  - Meetings (continued)
    - Regular Place (19)
    - Special (21)
- Board of Editors (46)
- Budget (26)
- Bylaws VI, 3
- Circuits Committee (46)
- Code of Administrative Practice (31)
- College Representatives (53)
- Committees
  - Appointment of VI, 7
  - Appointments (46), (49), (50)
  - Co-ordination by Executive Committee (24)
  - Direction by Executive Committee (24)
  - Personnel (46)
  - Standing (24) (46)
- Conferences (28)
- Constitution and Laws Committee (46)
  - Appointment (49)
- Constitution for Sections (40)
- Conventions (28)
- Co-operation by Sections (44)
- Copyright of Emblem (2)
- Credit, Sections (42)
- Debts, Sections (42)
- Delinquency, Sections (45)
- Diplomas, Fellow (13)
- Directors V, 1; VII, 2
  - Appointed V, 1
  - Elected V, 3
  - Vacancies VII, 3
- Discount on Subscriptions (56)
- Dissolution of Sections (45)
- Dues IV, 1
  - Arrears III, 7; (16)
  - Billing (14)
  - Mailing of Bills (15)
  - Notices (15)
  - PROCEEDINGS Mailing (9)
  - Subscription Allotment (56)
  - Waived IV, 3; (17)
  - When Payable IV, 2
- Duties of Committees (46)
- Editor
  - Appointment VII, 2; (49)
  - Executive Committee (51)
- Education Committee (46)
- Election of Officers VII
- Electroacoustics Committee (46)
- Electronics Committee (46)
- Emblem (2)
- Entrance Fee IV, 1
  - Payment for PROCEEDINGS (58)
- Establishment of Sections (38)
- Executive Committee (46)
  - Advertising (27)
  - Appointment (23)
  - Budget Preparation (26)
  - Code of Administrative Practice (31)
  - Committees
    - Appointment (50)
    - Co-ordination (24)
    - Direction (24)
  - Conferences (28)
  - Conventions (28)
  - Duties of Members (31)
  - Meetings (28)
  - Memberships
    - Approval (25)
    - Readmission (25)
  - Minutes of Meetings (32)
  - Office Management (27)
  - Overruled by Board (23)
  - Policies (33)
  - Publications (27)
  - Recommendations (33)
  - Salaries (27)
  - Sections (29)
  - Standards (30)
  - Suspend Power (23)
- Executive Management (23)
- Expulsions III, 8
- Facsimile Committee (46)
- Failure of Sections (45)
- Fellow
  - Abbreviations (1)
  - Citation (12)
  - Diplomas (13)
  - Dues IV, 1
  - Entrance Fee IV, 1
  - Nominees (12)
  - Privileges II, 1
  - PROCEEDINGS
    - Back Copies (58)
    - Subscription Rate (57)
  - Requirements II, 2
  - Symbol (1)
  - Transfer Fee IV, 1
- Financial Statement, Sections (43)
- Fiscal Year VI, 8
- Frequency Modulation Committee (46)
- Funds VI, 6
- Grade of Membership (1)
- Institute Representatives (53)
- Investment
  - Accounts (52)
  - Committee (24) (52)
- Investments (52)
- Joint Meetings, Sections (44)
- Laws and Constitution Committee (46)
- Mailing Address (15)
- Maintenance of Sections (41)
- Management VI
- Meetings VIII
  - Annual VIII, 1, 2
  - Board of Directors VIII, 3; (19)
    - Cancellation (20)
    - Place Change (20)
    - Regular (19)
    - Special (21)
    - Time Change (20)
  - Minutes of Executive Committee (32)
  - Sections
    - Joint (44)
    - Minimum (45)
    - Rebates (41)
    - Reports (43)
    - Requirements (45)
- Member
  - Abbreviation (1)
  - Dues IV, 1
  - Entrance Fee IV, 1
  - Newly Elected, Dues (6)
  - Newly Elected, Entrance Fee (6)
  - Posting of Names (5)
  - Privileges II, 1
  - PROCEEDINGS
    - Back Copies (58)
    - Mailing (7), (8)
    - Subscription Rate (57)
  - Reference (3)
  - Requirements II, 4
  - Symbol (1)
  - Transfer
    - Fee IV, 1
    - Membership Committee (11)
    - Others (10)
  - Use of Terms II, 11
- Membership
  - Admission (10)
  - Approval of Applications (25)
  - By Invitation (10)
  - Committee (46)
    - Personnel (47)
    - Section Chairman (47)
    - Transfer to Member (11)
    - Voided New Memberships (6)
  - Dues Allotment (57)
  - Grades II, 1
  - Readmission III, 6; (25)
  - Report of Secretary (18)
  - School Credit II, 9
  - Sections
    - Minimum (45)
    - Rebates (41)
    - Requirements (45)
  - Termination (14)
    - Membership Committee (14)
  - Transfer (10)
  - Use of Term II, 11

Minutes, Executive Committee (32)
Name I, 1
New Member
  Billing (6)
  Dues Payment Time (6)
  Notification III, 4
  PROCEEDINGS Mailing (7), (9)
  Voided Memberships (6)
Nominations VII
  Committee (35), (46)
    Appointment (49)
    Circularize Sections (34)
  Date of Submission (36)
  Multiple (35)
Notice of Election to Membership (6)
Object I, 2
Office Management (27)
Officers V, 1
  Appointment VII, 2
  Election VII, 1
  Nomination VII, 1; (36)
  Salary V, 5; (27)
  Terms V, 2, 4
Overrule of Executive Committee (23)
Papers Committee (46)
Petition
  Executive Committee Approval (37)
    For Sections (38)
Policies, Executive Committee (33)
Power, Executive Committee (23)
  Suspension (23)
President II, 1
  Duties VI, 1
  Executive Committee (51)
PROCEEDINGS
  Back Issues (16)
  Mailing
    Applicants (7), (8)
    Duration (9)
  Member Applicants (8)
  Name (55)
  New Members (7)
  Posting
    Elections (5)
    Transfers (5)
  Prior to Election (9)
  Senior Member Applicants (8)
  Rebate on Reinstatement (16)
  Subscriptions (56)
    Rate (57)
Publications (27)
  Receipt (14)
Public Relations Committee (46)
Radio Receivers Committee (46)
Radio Wave Propagation Committee (46)
Readmissions to Membership (25)
Rebates for Sections (41)
References (3)
  Associate Member (3)
  Senior Member (3)
  Student (3)
  Waiver (4)
Reinstatement III, 6
  PROCEEDINGS (16)
Requirements for Sections (45)
Representatives
  Institute (53)
  Other Bodies (54)
Resignations III, 5
Salaries (27)
  Secretary (22)
Secretary
  Appointment (49)
  Duties VI, 4
  Executive Committee (51)
  Investment Committee (52)
  Report on Membership (18)
  Salary (22)
Sections IX, (29)
  Authority, Limitation (42)
  Committees (46)
    Personnel (48)
    Sections Chairmen (48)
  Constitution (40)
Sections (continued)
  Co-operation with Others (44)
  Credit (42)
  Debts (42)
  Delinquency (45)
  Dissolution (45)
  Financial Statement (43)
  Formation IX, 1
  Joint Meetings (44)
  Maintenance (41)
  Meetings
    Joint (44)
    Minimum (45)
    Reports (43)
  Membership IX, 1
    Minimum (45)
  Petition (38)
  Rebates (41)
  Reports (43)
  Requirements (45)
  Termination IX, 1
  Territory (39)
Senior Member
  Abbreviation (1)
  Dues IV, 1
  Entrance Fee IV, 1
  Posting of Names (5)
  Privileges II, 1
  PROCEEDINGS
    Mailing (7), (8)
    Subscription Rate (57)
  Reference (3)
  Requirements II, 3
  Symbol (1)
  Transfer
    Fee IV, 1
    Membership (11)
Sponsors (10)
Standards (30)
  Committee (46)
Standing Committees (46)
  Appointment (50)
Student
  Dues IV, 1
  Entrance Fee IV, 1
  Privilege II, 1
  Relations (53)
  Representatives (53)
  Requirements II, 6
  School II, 8
  Reference (3)
Subscriptions to PROCEEDINGS (56)
  Discount (56)
  Rate (56), (57)
Symbols (1)
  Committee (46)
Television Committee (46)
Tellers Committee (46)
  Appointment (49)
Termination
  Membership (14)
  Sections (45)
Territory of Sections (39)
Transfer
  Associate
    Posting of Names (5)
  Fee IV, 1
  Member
    Posting of Names (5)
      Of Membership (10)
  Senior Member
    Posting of Names (5)
Transmitters Committee (46)
Treasurer V, 1
  Appointment (49)
  Duties VI, 5
  Executive Committee (51)
  Investments Committee (52)
Vice President V, 1
  Duties VI, 1
Voiding, New Memberships (6)
Voting II, 1, 12; (14)
Waiver of Dues at Age 65 (17)
Waiver of Sponsor (4)

Standards

on

RADIO WAVE PROPAGATION

DEFINITIONS OF TERMS
RELATING TO GUIDED WAVES

1945

THE INSTITUTE OF RADIO ENGINEERS

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by

The Institute of Radio Engineers, Incorporated

CONTENTS

Page
Introduction iv
Committee Personnel iv
Definitions of Terms 1
Index 3

INTRODUCTION

These definitions of terms were originally submitted to the Institute with the other terms on radio wave propagation which were published in 1942. They represent an agreement among the principal workers in this field that was obtained under the sponsorship of the Radio Wave Propagation Committee during 1940. These terms were carefully reviewed by the Radio Wave Propagation Committee during numerous meetings, by the Symbols Committee under whose responsibility these definitions were assigned by the Standards Committee when it selected them from the 1942 standards on radio wave propagation and again by the Radio Wave Propagation Committee upon reassignment to it of responsibility for them, and finally by the Standards Committee. They remain substantially the same as they were in 1940. Published with the approval of the Board of Directors, the report embodies the Institute's official recommendations to its members and the industry at large.

Suggestions and comments will be welcomed as an aid to committees preparing future reports. Correspondence should be addressed to The Institute of Radio Engineers.

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DEFINITIONS OF TERMS RELATING TO GUIDED WAVES

(Continued from Standards on Radio Wave Propagation—Definitions of Terms—1942)

SECTION 4. GUIDED WAVES

The waves that can be transmitted by cylindrical wave guides may be classified into two types: transverse electric (TE) for which the axial electric field is zero, and transverse magnetic (TM) for which the axial magnetic field is zero. If the line or lines which bound the cross section of the wave guide are lines for which one co-ordinate remains constant in a curvilinear co-ordinate system in which it is possible to separate the variables of the wave equation, the type of wave is designated by TE or TM followed by two subscripts indicating the appropriate solution in each of the two co-ordinates. The interpretation of the subscripts for wave guides of the more usual cross sections is given in the following definitions. Field distribution and cutoff frequencies are given for the more important types.

If the cross section of the wave guide does not fulfill the aforementioned requirement, the type of wave is designated by TE or TM with one numerical subscript starting with the wave that has the lowest frequency as "1" and numbering the remaining waves in the order of their cutoff frequencies.

4W1. Transverse Electric Wave (TE Wave). A transverse electric wave is any wave with the electric vector everywhere perpendicular to the direction of wave propagation.

4W2. Transverse Magnetic Wave (TM Wave). A transverse magnetic wave is any wave with the magnetic vector everywhere perpendicular to the direction of wave propagation.

4W3. Transverse Electromagnetic Wave (TEM Wave). A transverse electromagnetic wave is any wave with both the electric and magnetic vectors everywhere perpendicular to the direction of wave propagation.

4W4. Circular Magnetic Wave. A circular magnetic wave is any wave with circular magnetic lines of force.

4W5. Circular Electric Wave. A circular electric wave is any wave with circular electric lines of force.

4W6. Dominant Wave. A dominant wave is the guided wave which has the lowest cutoff frequency. It is the only wave which will carry energy when the excitation frequency is between the lowest cutoff and the next higher.

4W7. TE$_{n,m}$ Wave in a Circular Wave Guide. In a hollow circular metal cylinder the TE$_{0,m}$ wave is the circular electric wave (see Definition **4W5**) of order m; m is the number of cylinders, including the boundary of the guide on which the electric field vanishes. The TE$_{0,1}$ wave is the circular electric wave with the lowest cutoff frequency.

TE$_{n,m}$ waves ($n \neq 0$) are noncircular waves; n is the number of axial planes, and m, the number of cylinders including the boundary of the wave guide to which the electric vector is normal.

The dominant wave is the TE$_{1,1}$ wave, and its electric lines are approximately parallel to a diameter. The TE$_{0,1}$ wave is the circular electric wave with the lowest cutoff frequency.

In the TE$_{n,m}$ wave the axial magnetic field is proportional to $J_n(k_m\rho/a)$ cos $n\phi$ where n is the order of the Bessel function, k_m is the mth root of $J_n'(x)=0$, a is the radius of the cylinder, ρ is the distance from the axis, and ϕ is the angular co-ordinate.

4W8. TM$_{m,n}$ Wave in a Circular Wave Guide. In a circular metal cylinder the TM$_{0,m}$ wave is the circular magnetic wave (see Definition **4W4**) of order m; m is the number of cylinders including the boundary of the guide to which the electric vector is normal. The TM$_{0,1}$ wave is the circular magnetic wave with the lowest cutoff frequency.

TM$_{n,m}$ waves ($n \neq 0$) are noncircular waves; n is the number of axial planes to which the electric vector is normal, and m has the same meaning as above.

In the TM$_{n,m}$ wave the axial electric field is proportional to $J_n(k_m\rho/a)$ cos $n\phi$ where n is the order of the Bessel function, k_m is the nth root of $J_n(x)=0$, a is the radius of the cylinder, ρ is the distance from the axis, and ϕ is the angular co-ordinate.

4W9. TE$_{m,n}$ Wave in a Rectangular Wave Guide. In a rectangular metal cylinder the TE$_{m,n}$ wave is the transverse electric wave for which m is the number of half-period variations of the field along the x co-ordinate, which is assumed to coincide with the larger transverse dimension, and n is the number of half-period variations of the field along the y co-ordinate, which is assumed to coincide with the smaller transverse dimension.

The dominant wave in a rectangular wave guide is TE$_{1,0}$; its electric lines are parallel to the shorter side.

4W10. TM$_{m,n}$ Wave in a Rectangular Wave Guide. In a rectangular metal cylinder the TM$_{m,n}$ wave is the transverse magnetic wave for which m is the number of half-period variations of the field along the x co-ordinate, which is assumed to coincide with the larger transverse dimension, and n is the number of half-period variations of the field along the y co-ordinate, which is assumed to coincide with the smaller transverse dimension.

4W11. Resonant Modes in Cylindrical Cavities. When a metal cylinder is closed by two metal surfaces perpendicular to its axis a cylindrical cavity is formed. The resonant modes in this cavity are designated by adding a third subscript to indicate the number of half waves along the axis of the cavity. When the cavity is a rectangular parallelepiped the axis of the cylinder from which the cavity is assumed to be made should be designated since there are three possible cylinders out of which the parallelepiped may be made.

For TE$_{m,n}$ or TM$_{m,n}$ waves in hollow rectangular cylinders, the cutoff wavelength is $\lambda_c = 2/\sqrt{(m/a)^2+(n/b)^2}$ where a is the width of the guide along the x co-ordinate and b is the height of the guide along the y co-ordinate.

Table I gives the ratio of cutoff wavelengths to diameters for $TM_{n,m}$ waves in hollow circular metal cylinders:

TABLE I

		0	1	2	3	4	5 =n
	1	1.307	0.820	0.613	0.483	0.414	0.358
	2	0.569	0.448	0.373	0.322	0.284	0.2547
m =	3	0.363	0.309	0.270	0.241	0.219	0.200
	4	0.267	0.2375	0.2127	0.1938		
	5	0.2106	0.1910	0.1750			

Table II gives the ratio of cutoff wavelengths to diameters for $TE_{m,n}$ waves in hollow circular metal cylinders:

TABLE II

		0	1	2	3	4
	1	0.820	1.708	1.030	0.748	0.590
	2	0.448	0.59	0.468	0.382	0.338
m =	3	0.309	0.368	0.315	0.277	0.247
	4	0.2375	0.27	0.24	0.22	0.198
	5	0.1910	0.21	0.194	0.18	

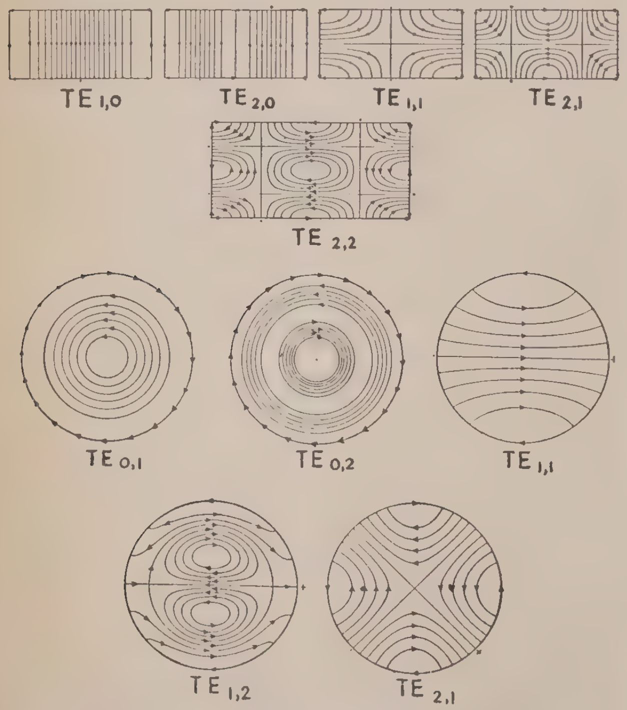

Fig. 1—The E lines for some of the more elementary TE modes in rectangular and circular wave guides.

The resonant wavelengths of cylindrical cavities are given by $\lambda_r = 1/\sqrt{(1/\lambda_c)^2 + (l/2c)^2}$ where λ_c is the cutoff wavelength for the transmission mode along the axis, l is the number of half-period variations of the field along the axis and c is the axial length of the cavity.

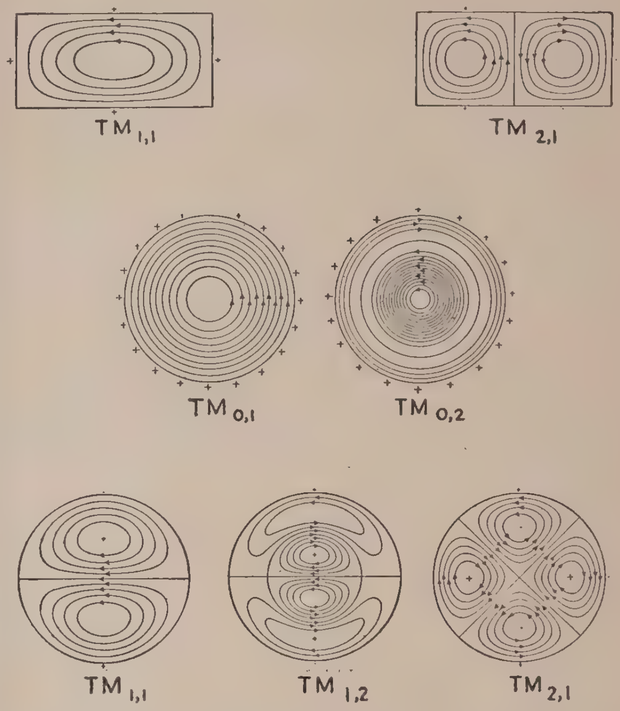

Fig. 2—The H lines for the more elementary TM modes of transmission in rectangular and circular wave guides.

ACKNOWLEDGMENT

The Committee wishes to acknowledge the kind permission of the publisher to reprint these figures from the book by S. A. Schelkunoff of the Bell Telephone Laboratories, entitled "Electromagnetic Waves," published by D. Van Nostrand Company, Inc., New York, N. Y.

INDEX

Definition Number

C

Circular Electric Wave . . . 4W5
Circular Magnetic Wave . . . 4W4
Cutoff Wavelengths in Circular Metal Cylinders . . . page 2
  $TM_{n,m}$
  $TE_{m,n}$

D

Dominant Wave . . . 4W6
Dominant Wave in a Circular Wave Guide . . . 4W7
Dominant Wave in a Rectangular Wave Guide . . . 4W9

R

Resonant Modes in Cylindrical Cavities . . . 4W11

T

$TE_{m,n}$ Wave in a Rectangular Wave Guide . . . 4W9
$TE_{0,m}$ Wave in a Circular Wave Guide . . . 4W7
$TE_{0,1}$ Wave in a Circular Wave Guide . . . 4W7
$TE_{1,1}$ Wave in a Circular Wave Guide . . . 4W7
$TE_{n,m}$ Wave in a Circular Wave Guide . . . 4W7
$TM_{m,n}$ Wave in a Rectangular Wave Guide . . . 4W10
$TM_{0,1}$ Wave in a Circular Wave Guide . . . 4W8
$TM_{n,m}$ Wave in a Circular Wave Guide . . . 4W8
$TM_{0,m}$ Wave in a Circular Wave Guide . . . 4W8
Transverse Electric Wave (TE Wave) . . . 4W1
Transverse Electromagnetic Wave (TEM Wave) . . . 4W3
Transverse Magnetic Wave (TM Wave) . . . 4W2

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on

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RECOMMENDED TERMINOLOGY

1945

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CONTENTS

Page

Introduction iv

Committee Personnel iv

Part I—General Recommendations 1

- 1. Definitions 1
  - Mother Crystal 1
  - Cut 1
  - Slab 1
  - Bar 1
  - Blank 1
  - Finished Blank, or Piezoid 1
  - Resonating Piezoid 1
  - Transducing Piezoid 1
- 2. Dimensions of Piezoid 1
- 3. Crystal Axes 1
- 4. Rotation of Axes 1
- 5. Orientation of a Plane, and Rotation about a Single Axis 2
- 6. Positive and Negative Direction of Rotation 2
- 7. Examples of Various Cuts, with Special Reference to Quartz 2
- 8. Stresses and Strains 3
- 9. Frequency Constant 3
- 10. Right- and Left-Handed Crystals 3

Part II—Special Conventions for Quartz Crystals 3

- 11. Axes of Quartz Crystals 3
- 12. The Two Forms of Quartz, and Methods of Distinguishing Them 4
  - (a) From the Natural Faces 4
  - (b) The Optical Method 4
  - (c) By X Rays 4
- 13. The *X* Axis and Piezoelectric Charges 4
- 14. Graphs of Quartz Crystals 4

Index 4

INTRODUCTION

This publication is a report of the Technical Committee on Piezoelectric Crystals. It contains definitions, conventions, and formulas, for the use of all persons who are concerned with piezoelectricity. It is directed especially to the attention of all authors in this field.

The Committee was first appointed in 1941 with the following membership: K. S. Van Dyke, *chairman;* S. A. Bokovoy, W. L. Bond, W. G. Cady, J. K. Clapp, W. P. Mason, and C. F. Baldwin. A few changes in personnel have taken place since then. At the time of issuance of this report the membership is as indicated below.

Besides holding meetings and carrying on much correspondence, the Committee has had the benefit of advice and suggestions from experts outside its membership. For all such help its sincere thanks are due.

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REPORT OF TECHNICAL COMMITTEE ON PIEZOELECTRIC CRYSTALS

This report is in two parts, the first containing general recommendations applicable to all piezoelectric crystals, and the second containing additional material that applies specifically to quartz.

PART I—GENERAL RECOMMENDATIONS

1. Definitions

1P1. Mother Crystal. The raw material.

1P2. Cut. A plane section with two parallel major surfaces cut in any orientation. Any given cut is specified by the direction normal to the major surfaces, or by special distinguishing symbols. Examples: an X-cut has its major surfaces normal to the X axis; in a Y'-cut the major surfaces are normal to the transformed Y' axis, etc. For other special designations see Part II.

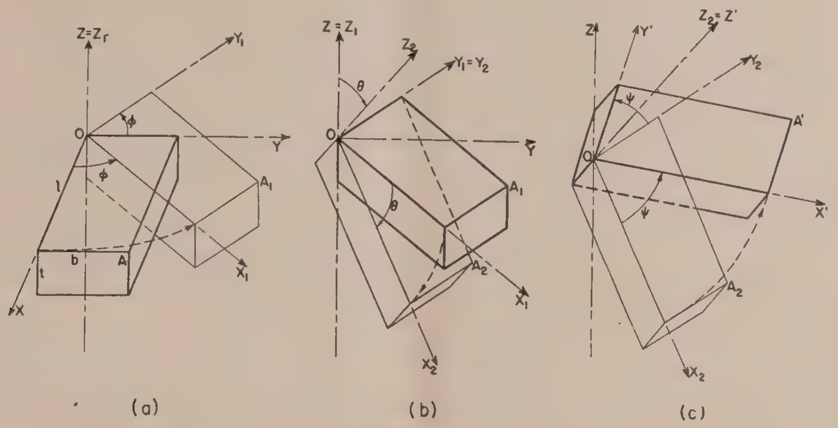


Fig. 1

(a) Rotation of a Z-cut piezoid through the azimuth angle ϕ.
(b) Second step: rotation through the polar angle Θ.
(c) Final step: rotation through the skew angle ψ.

As an aid in visualizing these diagrams one should note that in each rotation one edge of the parallelepiped remains fixed. In the rotation from OA to OA_1, the fixed edge is parallel to Z; it is parallel to Y_1 in the rotation from OA_1 to OA_2; and parallel to Z_2 in the rotation from OA_2 to OA'.

1P3. Slab. A relatively thick cut from which blanks are to be obtained by subsequent transverse cutting.

1P4. Bar. A subdivision of a slab, obtained by making two parallel saw-cuts in planes normal to the major surfaces of the slab.

1P5. Blank. The result of the final cutting operation.

1P6. Finished Blank or Piezoid.[1] The finished crystal product after the completion of all processes; this may include electrodes adherent to the crystal.

1P7. Resonating Piezoid. A piezoid used as a resonator or oscillator.

1P8. Transducing Piezoid. A piezoid used in a transducer.

2. Dimensions of Piezoid. When the piezoid is in the form of a rectangular parallelepiped, its length, breadth, and thickness should be denoted by l, b, (or w), and t, respectively. The only departure from this convention should be when there is danger of confusion, as for example when in the same discussion the symbol t is already being used for time or temperature. As an alternative, the symbol e (as used by the Curies and many later writers) is permissible.

3. Crystal Axes. A rectangular right-handed X, Y, Z system is always to be used, except in the case of enantiomorphic crystals of the *left* form, for which see Section 10. The X, Y, Z axes coincide with the crystallographic a_1, a_2, a_3 axes of the *cubic system;* with the a_1, a_2, c axes of the *tetragonal system;* with the a, b, c axes of the *rhombic system.* For the *hexagonal* and *trigonal systems* the three X axes coincide with the three Bravais axes a_1, a_2, a_3; the Z axis coincides with the Bravais c axis (optic axis); each Y axis is perpendicular to the Z axis and to an X axis. For the *monoclinic system* the X and Z axes coincide, respectively, with the c and b axes, and the Y axis is perpendicular to X and Z. The choice of axes for the *triclinic system* is arbitrary.

For the use of right- and left-handed axial systems see Section 10, and for the application to quartz see Section 11.

4. Rotation of Axes. For complete specification of the orientation of piezoids whose edges are oblique with

---

[1] The word "piezoid" should be pronounced thus: pie-ee'-zoid, with accent on the second syllable.

respect to all three crystal axes, a general transformation from the X, Y, Z axes to X', Y', Z' axes is required, involving three independent rotations, as follows:

Angle	*Symbol*
Azimuth angle	ϕ
Colatitude or polar angle	Θ
Angle of skew	ψ

The actual orientation of any piezoid is to be regarded as the result of turning it, by three successive operations, from an initial orientation corresponding to that of a Z-cut piezoid of which the l, b, and t dimensions are respectively parallel to the X, Y, and Z axes.

The first rotation is about the Z axis through the azimuth angle ϕ. This rotation is shown in Fig. 1 (a), in which the plate is rotated from the initial position OA to OA_1. The new axes are now X_1, Y_1, and $Z_1=Z$.

The second rotation consists in tilting the plate about the Y_1 axis through the polar angle Θ. The result is shown in Fig. 1 (b), the rotated position being OA_2 and the corresponding axes being X_2, $Y_2=Y_1$, and Z_2.

The final orientation of the piezoid is brought about by rotation about the Z_2 axis through the angle of skew ψ. The X_2 and Y_2 axes become X' and Y', and the final axial system is X', Y', and $Z'=Z_2$. The length, breadth, and thickness of the piezoid are now parallel respectively to X', Y', and Z', as is indicated by the parallelepiped OA' in Fig. 1 (c).

In specifying the orientation of piezoids, it is in general not necessary to trace the intermediate stages of rotation. The essential information is given by the values of ϕ, Θ, and ψ, together with the dimensions of the specimen.

Following is the matrix of direction cosines, for transforming from the X, Y, Z axes to the X', Y', Z' axes.

$$\begin{array}{llll} & X & Y & Z \\ X' & \cos\phi\cos\Theta\cos\psi - \sin\phi\sin\psi & \sin\phi\cos\Theta\cos\psi + \cos\phi\sin\psi & -\sin\Theta\cos\psi \\ Y' & -\cos\phi\cos\Theta\sin\psi - \sin\phi\cos\psi & -\sin\phi\cos\Theta\sin\psi + \cos\phi\cos\psi & \sin\Theta\sin\psi \\ Z' & \cos\phi\sin\Theta & \sin\phi\sin\Theta & \cos\Theta \end{array}$$

When a given angle can be stated in two ways, the numerically smaller value is to be used, except that 90 degrees is to be preferred to 30 degrees.

When there is a choice between a positive and a negative angle, the positive angle is to be used, unless it violates the foregoing rule.

5. ORIENTATION OF A PLANE, AND ROTATION ABOUT A SINGLE AXIS. When a *single direction* in the crystal is to be specified, it may be expressed either in terms of the direction cosines l, m, n, or in terms of the angles ϕ and Θ as defined above. In particular, the former method may be used to designate a *plane in any orientation*, in cases where it is not necessary to give the orientation of the edges of a piezoid in this plane.

In expressing the variation in such quantities as elastic and piezoelectric coefficients for rotation about a *single crystal axis*, it has been customary to designate the angle of rotation by θ, irrespective of the particular axis in question. It is recommended that this practice still be approved, provided that all particulars necessary to avoid confusion are stated.

6. POSITIVE AND NEGATIVE DIRECTIONS OF ROTATION. In any axial system, whether right- or left-handed, the positive direction of rotation is defined as follows:

Axis of Rotation	*Positive Direction of Rotation*
X	$+Y$ toward $+Z$
Y	$+Z$ toward $+X$
Z	$+X$ toward $+Y$

These rules apply also to systems with accented axes.

The following rules are equivalent to the above: in a *right-handed* system, a positive rotation is one that appears *counterclockwise* as viewed from the *positive end* of the axis of rotation. In a *left-handed* system, a positive rotation appears *clockwise* as viewed from the positive end of the axis of rotation.

7. EXAMPLES OF VARIOUS CUTS, WITH SPECIAL REFERENCE TO QUARTZ. Plates cut normal to a crystal axis may as well continue to be called X-cut, Y-cut, and Z-cut.

According to the system described above, the following well-known orientations of quartz piezoids are described thus:

	ϕ degrees	Θ degrees	ψ degrees
Z-cut piezoid	0	0	0
X-cut, length along Y	0	90	90
X-cut, length along Z	0	90	0
Y-cut, length along Z	30 or 90	90	0
Y-cut, length along X	90	90	90
X-cut bar, length in direction of maximum Young's modulus of quartz	0	90	48
AT-cut	−90	$54\frac{3}{4}$	90
BT-cut	90	41	90

The nature of the rotations is illustrated for certain special cases in Figs. 2 and 3. In Fig. 2 the initial position of the plate is shown at A. If $\phi=0$, $Y=Y_1$, and a rotation about Y_1 of $\Theta=90$ degrees brings the plate into position B. In this position we have an X-cut plate with length parallel to Z, and $\varphi=0$, $\Theta=90$ degrees, and $\psi=0$.

The group of quartz piezoids that have often been called Y'-cuts, and which now become Z'-cuts, can be derived from Fig. 3. In this figure the initial Z-cut plate has already been rotated through the azimuth angle $\phi=90$ degrees to position C. It is then turned through $\Theta=90$ degrees about the Y_1 axis to position D, in which position it is a Y-cut with length parallel to Z. If rotated further through a skew angle $\psi=90$ degrees about Z_2 it would have its length parallel to X. By letting Θ have other values, positive or negative, one obtains the group of tilted Y-cuts known as AT, BT, etc. In conformity with the notation of crystallographers, such a group as this, all of whose members have one dimension parallel to X, may be called an X group. This designation must not be confused with X-cuts.

8. Stresses and Strains. It is recommended that Voigt's symbol X_x, etc., be used for stresses, and x_x, etc., for strains. A compressional stress is by definition positive. A positive stress is associated with a negative strain.[2]

9. Frequency Constant. This term is defined as the number relating a natural vibrational frequency to a linear dimension of the piezoid. It is recommended that the symbol H be used for this quantity. If f is the frequency in kilocycles per second and d the linear dimension in millimeters, then

$$H = fd.$$

H is thus expressed in kilocycle-millimeters and should be abbreviated "kc-mm." It is numerically equal to the quotient (velocity of electromagnetic waves in meters per second)/(1000 h), where h is the familiar wave constant "meters per millimeter."

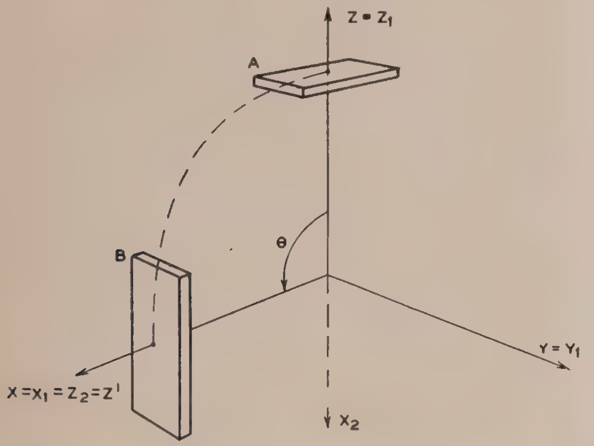


Fig. 2—An X-cut piezoid produced by rotation of a Z cut.

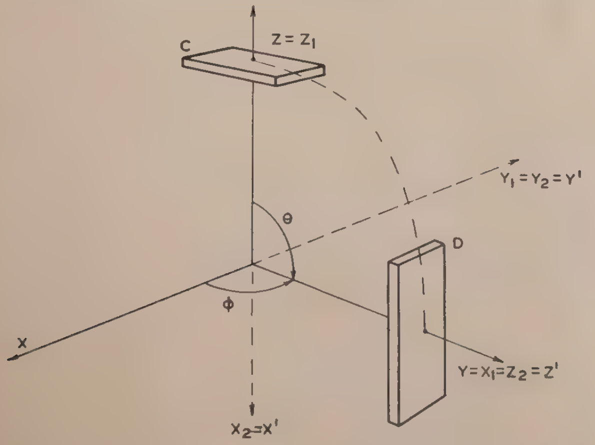


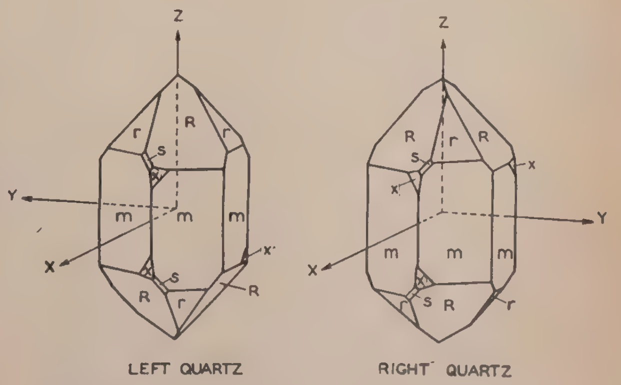
Fig. 3—A Y-cut piezoid produced by rotations of a Z cut.

10. Right- and Left-Handed Crystals. Crystals belonging to certain classes are enantiomorphic. That is, they occur in two different forms, in one of which the external faces and all internal properties are the mirror images of those in the other. All physical constants, diagrams, curves, and equations become identical for both forms when the axial system employed for a right (dextro) crystal is the mirror image of that for a left (laevo) crystal. We therefore recommend that a right-handed axial system be used for a right crystal, and a left-handed system for a left crystal. The corresponding rules for the sense of rotation for transformed axes have been given in Section 6.

Right and left quartz crystals are discussed in Part II. The only other crystal in common use belonging to an enantiomorphic class is Rochelle salt, and since it is found only in the dextro form, only a right-handed axial system is needed.

PART II—SPECIAL CONVENTIONS FOR QUARTZ CRYSTALS

11. The Axes of Quartz Crystals. The following definitions of axes apply equally to right and left quartz. On the orthogonal system, the Z *axis* is the optic axis, either end of which may be chosen as positive. As shown in Fig. 4, it is customary to represent the Z axis as pointing upward in the plane of the paper.



LEFT QUARTZ

RIGHT QUARTZ

Fig. 4—Right and left quartz crystals, showing prominent faces and axial systems.

Following are the Bravais symbols for the middle one of the three m faces in each diagram, for the s and x faces touching this m face at an upper corner, and for the R and r faces adjacent to this sx combination:

$$m = (10\bar{1}0) \qquad R = (1\bar{1}01)$$
$$r = (10\bar{1}1) \qquad s = (2\bar{1}\bar{1}1)$$
$$x = (6\bar{5}\bar{1}1)$$

The Bravais a_1 axis is taken as coinciding with the X axis shown in the figure, while the a_2 and a_3 axes coincide with the other two X axes, taken in counterclockwise order (as seen from the positive end of the Z axis) for right quartz, clockwise for left quartz. By this convention corresponding faces have the same symbols for both forms of quartz.

---

[2] A discussion of the various types of strain is given by W. G. Cady and K. S. Van Dyke in "Proposed standard conventions for expressing the elastic and piezoelectric properties of right- and left-hand quartz," Proc. I.R.E., vol. 30, pp. 495–499; November, 1942.

There are three equivalent X or electrical axes, and correspondingly three equivalent Y or mechanical axes. Each *X axis* bisects the angle between two prismatic m faces, its positive end projecting outward at an edge where, in the ideal crystal, the *sx* combinations are found.

Each *Y axis* is normal to a pair of opposite prismatic m faces. The positive end of each points outward from a prismatic face that intersects a major rhombohedral R face at the $+Z$ end.[3]

The axes as thus defined form a right-handed system for right quartz, left-handed for left quartz.

12. THE TWO FORMS OF QUARTZ. Left quartz can be distinguished from right in several different ways, three of which are as follows:

(a) *From the Natural Faces.* The crystal may show one or more of the so-called x faces, or better still one or more of the *sx* combinations. This combination is shown in Fig. 4. The form at the left is the mirror image of that at the right with respect to the ZX plane. The figure shows that when a quartz crystal is held with the optic (Z) axis vertical and with an *sx* combination toward the front and at the upper end of a prismatic edge, then in a right quartz the normal to the intersection of these two faces points upward and to the right, while in a left quartz it points upward and to the left. The nonparallel edges of an x face converge upward toward the right in a right quartz, upward toward the left in a left quartz.

(b) *The Optical Method.* Since quartz is optically active, the plane of polarization in a beam of plane-polarized light parallel to the optic axis becomes rotated about this axis. In conformity with what seems to be the most widely accepted practice, we recommend that a quartz crystal be called *right* when on examination in a polariscope it rotates the plane of polarization in such a way that the analyzer, if viewed toward the source of light, must be rotated clockwise to follow the rotation. If the analyzer has to be rotated counterclockwise, the specimen is a left quartz.

(c) *By X Rays.* In the YZ plane, 14° 42′ above the Y axis (between $+Y$ and $+Z$), there is a normal to the atomic plane $(03\bar{3}1)$ that reflects X rays. (The plane whose normal is 14° 42′ below Y does not reflect.) The observed X-ray reflection determines $+Y$ and $+Z$. The positive direction of the X axis is found by piezoelectric tests (see Section 13). The specimen is of right or left quartz according to whether X, Y, and Z are found to form a right- or left-handed system.

Methods (a), (b), and (c) always give identical results in distinguishing between the two forms of quartz.

13. THE X AXIS AND PIEZOELECTRIC CHARGES. It is a recognized fact that the prismatic edge below an s face becomes negatively charged by compression along the X axis that emerges at this edge. Since a compression produces a negative strain and an extension a positive strain, it follows, according to the present recommendation concerning axes, as illustrated in Fig. 4, that the positive direction of an X axis can be defined by the following rule: The positive end of an X axis is that end at which a positive charge appears when the strain along this axis is positive.

14. GRAPHS OF QUARTZ CRYSTALS. It is suggested that all graphical representations of quartz crystals in published papers be drawn for *right* quartz, hence with a right-handed system of co-ordinate axes, except when there is some special reason for doing otherwise.

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[3] It is recommended that the major and minor rhombohedral faces be called the R and r faces.

INDEX

Page
Axes . . . . . . 1§3, 1§4, 2§5, 3§11, 4§13
Bar . . . . . . 1§1
Blank . . . . . . 1§1
Cut . . . . . . 1§1, 2§7
Definitions . . . . . . 1§1
Direction in Crystal, Specification of . . . 2§5
Enantiomorphic Crystals . . . . 3§10,
Frequency Constant . . . . . . 3§9
Graphs of Quartz Crystals . . . . 4§14
Left-Handed System . . . . . . 2§6, 3§10
Orientation of Piezoid . . . . . . 1§4, 2§7
Piezoelectric Charges on Quartz . . . . 4§13
Piezoid . . . . . . 1§§1, 2
Plane, Orientation of . . . . . . 2§5
Quartz . . . . . . 2§7, 3§11, 4§§12, 13
Right- and Left-Handed Crystals . . . . 3§10, 4§12
Rotation of Axes . . . . . . 1§4, 2§§5, 6
Slab . . . . . . 1§1
Stresses and Strains . . . . . . 2§8
Symbols Recommended . . . . . . 1§§2, 4
Tests for Right and Left Quartz . . . . 4§12
Wave Constant . . . . . . 3§9

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VOLUME 33 — *December, 1945* — NUMBER 12, PART I

Section Meetings 820
I.R.E. Members, Their Institute, and Their PROCEEDINGS. The Editor 821
William O. Swinyard—Board of Directors—1945 822
The Engineer's Place in Naval Research .... W. G. Schindler 823
A Review of Plastic Materials .... H. L. Brouse 825
Effect of Surface Finish and Wall Thickness on the Operating Temperature of Graphite Radio-Tube Anodes .... L. L. Winter and H. G. MacPherson 834
A Coil-Neutralized Vacuum-Tube Amplifier at Very-High Frequencies .... R. J. Kircher 838
Cathode-Follower Circuits .... Kurt Schlesinger 843
Television Resolution as a Function of Line Structures. Madison Cawein 855
Loop-Antenna Coupling-Transformer Design .... W. S. Bachman 865
On the Theory of the Progressive Universal Winding .. Alfred W. Simon 868
Concerning Hallén's Integral Equation for Cylindrical Antennas .... S. A. Schelkunoff 872
Frequency-Modulation Distortion Caused by Multipath Transmission .... Murlan S. Corrington 878
Symmetrical Antenna Arrays .... Charles W. Harrison, Jr. 892
General Formulas for "T"- and "Π"-Network Equivalents. Myril B. Reed 897
Institute News and Radio Notes 900
- Plans Go Forward for Winter Technical Meeting and Radio Engineering Show 900
- Board of Directors 900
- Executive Committee 901
- Browder J. Thompson Memorial 902
- I.R.E. People 902

Books:

- "The Decibel Notation and Its Applications to Radio Engineering and Acoustics," by V. V. L. Rao .... R. V. L. Hartley 903
- "On the Propagation of Radio Waves," by Olof E. H. Rydbeck. .... Kenneth A. Norton 903
- UHF Radio Simplified, by Milton S. Kiver .... O. S. Duffendack 903

Correspondence:

- "Note on Iron Powder Cores and Coils". H. W. Jaderholm 904

Contributors 904
Section Meetings 34A
Membership 38A
Positions Open 50A
Advertising Index 86A

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BALTIMORE	—	R. N. Harmon	F. W. Fischer	714 S. Beechfield Ave., Baltimore, Md.
BOSTON	—	C. C. Harris	A. G. Bousquet	General Radio Co., Cambridge 39, Mass.
BUENOS AIRES (Argentina)	—	A. DiMarco	H. Krahenbuhl	Transradio Internacional, San Martin 379, Buenos Aires, Argentina
BUFFALO-NIAGARA	December 19	J. M. Van Baalen	H. W. Staderman	264 Loring Ave., Buffalo, N. Y.
CEDAR RAPIDS	—	F. M. Davis	J. A. Green	Collins Radio Co., 855—35 St., N.E., Cedar Rapids, Iowa
CHICAGO	December 21	Cullen Moore	L. E. Packard	General Radio Co., 920 S. Michigan Ave. Chicago 5, Ill.
CINCINNATI	December 18	L. M. Clement	J. F. Jordan	The Baldwin Co., 1801 Gilbert Ave., Cincinnati 2, Ohio
CLEVELAND	December 27	H. B. Okeson	A. J. Kres	16911 Valleyview Ave., Cleveland 11, Ohio
CONNECTICUT VALLEY	December 20	H. W. Sundius	L. A. Reilly	989 Roosevelt Ave., Springfield, Mass.
DALLAS-FORT WORTH	—	J. D. Mathis	B. B. Honeycutt	9025 Roanoak, Dallas 18, Texas
DAYTON	December 20	L. B. Hallman	Joseph General	411 E. Bruce Ave., Dayton 5, Ohio
DETROIT	December 21	L. H. Larime	R. R. Barnes	1810 Sycamore, Royal Oak, Mich.
EMPORIUM	—	W. A. Dickinson	H. E. Ackman	West Creek, R. D. 2, Emporium, Pa.
INDIANAPOLIS	—	H. I. Metz	E. E. Alden	4225 Guilford Ave., Indianapolis, Ind.
KANSAS CITY	—	R. N. White	Mrs. G. L. Curtis	6003 El Monte, Mission, Kan.
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LOS ANGELES	December 18	R. C. Moody	R. G. Denechaud	Blue Network Co., 6285 Sunset Blvd., Hollywood 28, Calif.
MONTREAL (Canada)	January 9	L. A. W. East	R. R. Desaulniers	Canadian Marconi Co., 211 St. Sacrament St., Montreal, Que., Canada
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OTTAWA (CANADA)	December 20	W. A. Steel	L. F. Millett	33 Regent St., Ottawa, Ont., Canada
PHILADELPHIA	—	D. B. Smith	P. M. Craig	Philco Corp., Philadelphia 34, Pa.
PITTSBURGH	December 10	J. A. Hutcheson	C. W. Gilbert	52 Hathaway Ct., R.F.D. 1, Wilkinsburg 21, Pa.
PORTLAND	—	Kenneth Johnson	C. W. Lund	Route 4, Box 858, Portland, Ore.
ROCHESTER	December 20	G. R. Town	A. E. Newlon	Stromberg-Carlson Co., Rochester 3, N.Y.
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SAN FRANCISCO	—	David Packard	William Barclay	955 N. California Ave., Palo Alto, Calif.
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PRINCETON	—	W. C. Johnson	J. G. Barry	Princeton University, Princeton, N. J.
SOUTH BEND	December 20	H. E. Ellithorn	J. E. Willson	1002 S. Lombardy Dr., South Bend, Ind.

I. R. E. Members, Their Institute, and Their PROCEEDINGS

It may be stated with propriety that the membership of The Institute of Radio Engineers includes a thoroughly competent group of communication and electronic engineers in the active practice of their profession at this time. The Institute is the agency through which their collective activities are stimulated and their professional viewpoints collected, co-ordinated, and disseminated. The PROCEEDINGS of the I.R.E. is the chosen instrument whereby the work of each Institute member reaches his professional associates. And as a result of the activities of the membership of the Institute, and of the publication of their PROCEEDINGS, the electronic and communication arts and sciences are notably advanced. Applications of engineering knowledge to the benefit of mankind are increased in number and importance. The industries based upon the devices and techniques developed by the I.R.E. membership are stimulated in their growth and led to greater usefulness and prosperity. And the engineering members of the Institute are themselves benefitted and encouraged by the progress and expansion of their art and the increasing public esteem in which they are held.

The founders of the Institute and the long line of competent and devoted executives, Section officers, Committee Chairmen, Committee members, and Institute employees have contributed in great measure to the growth of the Institute, not only numerically or quantitatively but also in the quality and merit of its contributions. In the last analysis, nevertheless, it is the membership of the Institute who in the past have made the Institute possible and who in the future must continue to maintain and enhance its standing. The solidarity of the Institute membership with their professional society is in fact the keystone not only of future I.R.E. accomplishments but also of the standing and prosperity of the engineers who are its members and of the industries in which they work.

It may be gladly acknowledged that the industries in question have demonstrated their understanding of the constructive and mutually beneficial relationship between themselves and the Institute. The Building-Fund campaign has shown that most of the industrial leaders already understood the purposes and accomplishments of the Institute. And those who at first did not fully appreciate its worth were soon convinced and became, like their colleagues, its friends and supporters.

It is not only inevitable but desirable that the great expansion in the membership and activities of the Institute should require the expansion of its employed staff through the addition of skilled individuals who are capable of carrying, year in and year out, many of the burdens of administration and operation. Policy decisions of major professional import should and will remain in the hands of the engineering executive officers and directors of the Institute, selected to that end by the Institute membership. But daily operations and the corresponding decisions can no longer be handled by active professional engineers engrossed in their own technical activities and able to devote to the Institute no more than limited periods of time at certain intervals. The Institute requires continuous supervision and operation, and the burden of such operation is no longer a part-time task.

The Institute membership and directors are well aware of these conditions and have accordingly understood and approved the necessary staff expansions, reorganization, and improved operational methods which have been introduced. Such officers as the Executive Secretary, the Assistant Secretary, and the Technical Secretary, in the Secretarial Department as well as the Associate Editor and the Technical Editor, in the Editorial Department will enable the more efficient and productive handling of the Institute's ever-increasing activities.

It is the obligation of these members of the Institute staff to plan and execute their duties dependably and in the interests of the membership. They merit the support and friendliness of the Institute membership. It is the duty and opportunity of the officers and directors of the Institute to determine the broad policies of the Institute and to guide the employees of the Institute along such major lines as will ensure the advancement of the communication and electronic engineering profession. And it is the privilege and opportunity of the membership of the Institute to aid and guide both of these groups; to express opinions of commendation or criticism freely; to select the policy-forming group of the Institute; to participate in the Committee, Convention, Section, and publication activities of the Institute; and to feel themselves ever more closely associated through a bond of common membership, friendliness, and ideals in one of the great professional organizations of the world of our times.

The Editor

William O. Swinyard

Board of Directors—1945

William O. Swinyard was born on July 17, 1904, at Logan, Utah. He received the B.S. degree in mathematics and physics in 1927 from the Utah State College. From 1927 to 1930 he was an instructor in physics, mathematics, and music in secondary schools in southern Idaho. In the fall of 1930 he was employed in the Bayside, Long Island, laboratory of the Hazeltine Electronics Corporation. During the next four years Mr. Swinyard worked on general engineering problems and later was actively engaged in some of the early work done by the company on high-fidelity broadcast receivers and associated measuring equipment. In 1934 he was transferred to Hazeltine's New York laboratory where he spent three years as a senior engineer working on the design and development of commercial broadcast receivers and the measurement of their performance characteristics. In 1937 he was transferred to the Chicago laboratory, as assistant engineer in charge, and has been engineer in charge of that laboratory since 1942.

He is the author of several technical papers and book reviews which have been published in the PROCEEDINGS and in other technical magazines.

Mr. Swinyard joined the Institute of Radio Engineers as an Associate in 1937 and was transferred to the Senior Member grade in 1939. He has been active in I.R.E. section activities since 1937. At present, he is Chairman of the Chicago Section and a member of the I.R.E. Education Committee and Sections Committee.

Mr. Swinyard is a charter member of the Illinois professional Communications Engineers Association and was chairman from 1942–1944. He is president of the Radio Engineers Club of Chicago. He was a member of the National Electronics Conference Executive Committee in 1944, and is chairman-elect of the Executive Committee for 1945.

Mr. Swinyard also is a member of the Radio Club of America, the Acoustical Society of America, and the Chicago Physics Club.

The Engineer's Place in Naval Research*

WALTER G. SCHINDLER†

IN THE May, 1945, issue of the PROCEEDINGS OF THE I.R.E., a very interesting symposium entitled "The Engineer's Place in the Scheme of Things" is reviewed in brief. Many engineers of Naval Ordnance are in sympathy and agreement with many of the views expressed in the article. There were, however, some remarks appearing as paragraph three, column one, of page 288, with which we of the Naval Ordnance Laboratory wish to take exception. In justice to what might be called the government defense-research services, since the beginning of the national emergency and certainly in the future, it is believed that the impressions created by the paragraph should be corrected.

The paragraph states: "Another interesting matter considered by him was the comparison of government work with industrial work, and correlatively the likely future of an engineer in a government organization. The danger of government-laboratory work, he felt, is that future important developments might never spread out into private industry and would thus be immured in the governmentally organized laboratories. Industry can envisage better peacetime applications of research than can government organizations. Competent men, he feared, might be taken from private industries and placed in government work which would not advance the development of industry and engineering as effectively as if they were in private industries. Thus many engineers might be frozen in their present status and not serve to advance peaceful progress. A government laboratory engineer cannot, in general, be industrially active after the war unless he goes into industry."

In refutation it should first be pointed out that national-defense research and development is so strikingly similar to industrial research and development that much of the force of the statement is lost. Like industrial research, naval research derives from the fruits of the fundamental research of the pure scientist. It starts with basic studies of new fundamental research phenomena with an eye to application. It proceeds through the creative or inventive phase of evolving a device to achieve a technical objective. The device passes through the laboratory or bread-board-model stage, through the design and development phase of an adequately tested working model, through pilot production and mass production stages to introduction into the field by means of technical salesmanship. Like industrial research it is developed under the same conditions of secrecy, or security as it is termed by the military. It is likewise protected by patents in the same fashion as the industrial devices. There are only two respects in which the device developed by the Navy or the government defense agency differs from the device developed by industry. One of these is the degree of secrecy or security maintained. The other lies in the character of the distribution, employment, and the type of technical salesmanship involved in the later use of the device. As regards the security aspect, it must be stated that even this, as in industry, enjoys a wide scope of security levels. These range all the way from free dissemination through the classifications termed "restricted," "confidential," "secret," and "top secret," depending on the nature of the device and its uses. The industrial devices until safely protected by patents, are usually kept at the confidential or secret level, and no taxpayers' representatives can pry into these secrets. Thereafter, when the sales phase demands publicity, the industrial-research achievement and the device are given wide publicity. Even then, there are certain concealments and reticences referred to as "trade secrets" which prevent complete revelation of all details. While such publicity will not accompany very many of the naval devices, there are relatively few devices in the confidential, secret, or top secret brackets. It is remarkable to what extent even the more highly classified devices leak out into technical journals and magazines. In fact the sustained higher classification applies especially to details of a few specially designed weapons or devices of a critical nature. Needless to say, in time of war the maintenance of security is paramount for national welfare, and therefore it is improper reasoning to project the present security measures into the peacetime future.

As regards the technical salesmanship phase of industry, this is duplicated in naval and military spheres in the accompanying of the device into the fleet or the field by groups of qualified engineers whose function it is to sell the device to the services, to service it in technical employment, and to evaluate and exploit its tactical potentialities. In the armed services this phase is, however, on a classified level and is available only to the services and co-operating engineers from industry.

It should, in passing, also be noted that in naval and military research and development industrial concerns are in a large, in fact preponderating measure, the big brothers or partners of the armed services. It is they who in an overwhelming measure produce the articles and weapons of war in our common defense endeavor. Industry is therefore as much involved in these classified items as are the research laboratories of the armed services. It thus makes little difference whether the engineers are employed by industry for the work on these items or by the armed services. This close co-operation and common interest must be emphasized in answer to the criticism.

Before proceeding further, it is essential to stress the vital importance of technical defense research after the war. For the preservation of our country, our way of life, and of our industries, it is imperative that, in the uncertain peace of the future, we adequately prepare in time of peace for the new weapons of the future and the countermeasures thereto. In the measure that we are prepared, in that measure will we discourage attack by a potential enemy, and in that measure only will we be able to survive. Adequate peacetime national-defense research for the future is as urgent as such research in time of war. It is as much the concern of industry as it is of the engineer, the scientist, and the government defense-research agencies. It behooves our engineering

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† Naval Ordnance Laboratory, Washington 25, D.C.

profession in time of peace, as in time of war, to assist and encourage such work, not to discourage it.

The danger that such research laboratories will make serious inroads on industry by taking too many good engineers, as indicated in the paragraph cited, is preposterous for two good reasons. The first is that in the imminent peacetime economy there will be all too few of such agencies. The second reason is that the paragraph cited indicates ignorance of the plans which are contemplated for future government defense agencies, certain portions of which are already in use for the conduct of such research. It is believed that, when such plans are understood, readers will agree that the technical defense-research laboratories will not be the evil predicted, but will do a positive service for the engineering profession.

The agencies which will carry on these researches are:

(a) Research laboratories of the Army and the Navy. There must be such laboratories to carry on certain problems directly connected with weapons, their development, and improvement because of their ability to work closely with the armed services. They will supposedly also handle the more highly classified weapons.

(b) Research units operating under the cognizance of the Research Board for National Security in co-operation with the armed forces. These units will be few and will work in industrial or academic research laboratories under contract for specific projects of limited duration.

(c) Industrial research laboratories engaged in working on devices of interest to the armed forces, or devices which will be of service in national defense.

As regards the conduct of the research laboratories operating under the armed services, the following comments should be made. Before World War II these laboratories were painfully inadequate as to staff, facilities, and funds, and fell heir to the evils resulting from lack of appreciation and support. With the establishment of the national emergency, early in 1940, these laboratories were rapidly and vastly expanded. New ones were established. Research units were set up under the Office of Scientific Research and Development. Contract employment together with the patriotism of our American scientists and engineers, resulted in the employment of the ablest men in this country for work on these problems. Under the stimulus and guidance of these men, what seemed an impossible task of technical achievement for national defense was accomplished in time to bring victory. In the next war the required time may not exist. In this war it was accomplished, but barely in time. These men are still with us. Many, it is hoped, will remain with us after the war. They have seen the difficulties of the past. They plan to avoid them in the future. These plans envisage the following procedures:

(a) To establish such means of hiring by contract and under Civil Service that proper recruiting and selection of personnel comparable to those currently used by industry will be possible.

(b) To establish adequate pay grades comparable with industry.

(c) To establish classification and promotion systems such that employees may be rated for technical efficiency as well as for administrative efficiency.

(d) To permit of liberal in-grade increase of salaries and to liberalize promotion procedures.

(e) To encourage membership of its technical personnel in technical societies, and to encourage participation in scientific meetings.

(f) To encourage, as far as possible, sound basic research and to permit very much wider publication of results of purely scientific interest. Despite the war conditions still existing, the Naval Ordnance Laboratory has, in the last six months, submitted in excess of a dozen papers for publication in scientific ournals. Once the pressure of wartime production is eased, many more papers of considerable engineering value will be published.

(g) To encourage and achieve exchange of technical personnel for periods of the order of months to a year among industrial laboratories, scientific laboratories, and the defense-research laboratories.

(h) To facilitate advanced study of its junior personnel and to have scientific seminars with outside and local speakers.

(i) To organize a technical society on a classified level for all defense-research agencies.

(j) To arrange for work or study by its personnel at appropriate industrial and educational research laboratories.

(k) Not to encourage its personnel to remain too long in work of this character. That is, to hire able, brilliant, and capable young technical men, to train them, advance them as rapidly as possible, and to encourage them to seek outside employment at suitable times after, perhaps, intervals not to exceed ten years.

(l) To place the direction of technical research in the hands of competent technical men, and to reduce to a necessary minimum the administrative supervision of the nontechnical military personnel. On the other hand, to co-operate with them in the achievement of common objectives.

(m) To organize the laboratories in terms of stimulating scientific and technical groups, and to exchange duties within the laboratory, insofar as is consistent with good judgment, to broaden the background and experience of the personnel.

(n) To furnish the best of equipment and material facilities, including libraries, to its personnel.

(o) To encourage originality, genius, and enterprise among the technical employees to the utmost compatible with the accomplishment of the tasks in hand.

It is clear that, if these objectives can be achieved, the engineer engaged in technical defense research for the government will be better off than if employed in many smaller industries, and that in that measure the engineering profession will be benefited. In some directions indicated by this program, the government defense-research laboratories have already achieved material success. There is much more to be done in the future. With the understanding, endorsement, and support of the engieering profession, we stand a good chance of achieving complete success. We hope that we can count on that co-operation.

A Review of Plastic Materials*

HAROLD L. BROUSE†, MEMBER, I.R.E.

***Summary*—The spectacular growth of the plastics industry to s present state is a direct result of proved ability to meet war de-ands for critical items on a volume-production basis.**

Common plastic materials are classified according to their ther-al characteristics and the methods of fabrication.

Typical physical and electrical properties are presented for each f the general class of materials. It is pointed out that such values re average, and do not indicate the range that can be obtained by election of a specific molding powder.

Common trade names, compositions, and their manufacturers are resented in the form of a cross-index for reference purposes.

Factors governing the performance and the choice of a particular aterial are discussed.

Expansion coefficients as applied to the use of metal inserts olded into the material, light transmission properties, and the brication of laminates are briefly treated in the Appendix.

INTRODUCTION

TREMENDOUS advances have been made during the past several years in all phases of the plastics industry. Briefly, new and improved molding owders have been developed by the basic-materials anufacturers; press equipment has been redesigned so s to afford better control of molding temperatures and ressures; radio-frequency dielectric heating has become n accepted practice; and new techniques in the art of nolding have been conceived. Hence, today's produc-ion of plastics items for the armed forces typifies the fficient teamwork "behind the scenes" necessary to upply the best-equipped army in the world.

Plastics may be looked upon as an infant member of he large organic family; not new-born, but in a dy-amic state of development today. In order to evaluate he qualities of new materials as they become available, ne should be able to associate their properties with the haracteristics of similar, older materials known from ast experience.

CLASSIFICATION OF MATERIALS

While there is a variety of ways in which materials an be classified, the one most generally used is accord-ng to whether the material is hardened upon the ap-lication of heat; i.e., thermosetting, or whether it can e softened by reheating, viz., thermoplastic. Table I ists the principal types of materials in each of these two groups. Tables X and XI provide a cross-index of manu-acturers, trade names, and compositions.

TABLE I

PRINCIPAL THERMOSETTING MATERIALS

- Allyl Resins
- Aniline Formaldehyde
- Casein
- Lignin
- Melamine Formaldehyde
- Phenol Formaldehyde
- Phenol Furfural
- Urea Formaldehyde

PRINCIPAL THERMOPLASTIC MATERIALS

- Amides (Nylon)
- Cellulose Resins
  - Cellulose Acetate
  - Cellulose Acetate-Butyrate
  - Cellulose Nitrate
  - Ethyl Cellulose
- Methyl Methacrylate
- Polyethylene
- Polyvinyl Alcohol
- Shellac
- Silicones
- Styrene Resins
  - Polystyrene
  - Polystyrene base (Styramic)
  - Polydichlorostyrene (Styramic HT)
  - Styrene Copolymers
    - Cerax
    - Styraloy
- Vinyl Aldehyde Resins
  - Polyvinyl Acetal
  - Polyvinyl Butyral
  - Polyvinyl Formal
- Vinyl Ester Resins
  - Polyvinyl Acetate
  - Polyvinyl Chloride
  - Polyvinyl Chloride-Acetate
- Vinylidene Chloride (Saran)

COLD-MOLDING COMPOUNDS

Cold-molding compounds comprise a group of ma-erials which have been used extensively for more than wo decades. As indicated by their name, the powder is cold-pressed to shape in suitable dies, after which the "green" piece is subjected to an oven-baking operation. Owing to the high content of asbestos filler used in the compounds, the molded parts are extremely resistant to heat, and are widely used for such applications as connector plugs and handles, arc shields, and electrical heating-element supports. These compounds may be classified as shown in Table II.

TABLE II

COLD-MOLDING COMPOUNDS

1. Inorganic binder (refractory type)
   a. Lime-silica cement base

2. Organic binder (nonrefractory type)
   a. Bituminous base
   b. Synthetic-resin base (phenolic, etc.)

FILLERS AND BASE MATERIALS

In most instances, a molding powder will contain a number of ingredients in addition to the resin itself, blended so as to facilitate molding and to impart desirable characteristics in the molded piece. Such compounds consist of fillers or extenders, catalysts, solvents, lubricants, plasticisers, etc. A filler may be used to obtain added impact strength, improved electrical properties,

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† The Crosley Corporation, Cincinnati, Ohio.

or added heat resistance, depending upon the particular characteristic desired. Fillers become extenders when they are added merely to reduce the cost of the molding material by replacing a portion of the resin with a cheaper, inert material. On the other hand, laminates consist of a base material which is built up using a properly blended natural or synthetic resin as a binder and/or impregnant. Typical fillers used in molding powders and base materials used for laminates are indicated in Table III.

TABLE III

FILLERS AND BASE MATERIALS

Molding Compound Fillers

- Alpha Cellulose
- Asbestos
- China Clay
- Cotton Flock
- Cotton Fabric
- Graphite
- Linen Fabric
- Mica
- Sisal Fiber
- Silica
- Slate Dust
- Tire Cord
- Wood Flour

Laminated Base Materials

- Asbestos
- Cotton Fabric
- Fiber-Glass
- Linen
- Macerates
- Paper
- Wood Pulp
- Wood Veneers

METHODS OF FABRICATION

Progress in all phases of the industry has made possible new and higher productive methods of fabrication. Table IV shows the methods in use at the present time for handling thermosetting and thermoplastic materials.

TABLE IV

METHODS OF FABRICATION

- Thermosetting Materials
  - Casting
  - Compression Molding
  - Jet Molding
  - Laminates
    - High pressure, steam cure
    - High pressure, radio-frequency cure
    - Low pressure, bag molding
    - Contact pressure
  - Macerates and Pulp Molding
  - Post Forming of Laminates
  - Transfer Molding
- Thermoplastic Materials
  - Blowing
  - Calendering
  - Compression Molding
  - Deep Drawing
  - Extrusion Molding
  - Film Casting
  - Injection Molding
  - Post Forming
  - Pressing

Naturally, all of the fabrication methods are not used for any single material since these methods cover those processes which have been developed to meet specific needs. However, certain groupings are evident; e.g., thermosetting materials which are compression molded can, likewise, be transfer molded or jet molded, provided the filler is not too bulky. By the same token, most thermoplastic materials, which can be injection molded, can be extruded.

The more common methods of fabrication applicable to each of the plastics groups are presented in Table V.

TABLE V

FABRICATION METHODS APPLICABLE TO VARIOUS PLASTIC MATERIALS

Thermosetting Materials	Casting	Impregnating Resins	Compression Molding	Injection[1]	Extrusion	Calendering	Blowing	Pressing	Film
Allyl Resins	X	X							
Aniline Formaldehyde			X						
Casein			X					X	
Lignin		X	X						
Melamine Formaldehyde		X	X						
Phenol Formaldehyde	X	X	X						
Phenol Furfural	X	X	X						
Silicones		X							
Urea Formaldehyde		X	X						
Thermoplastic Materials									
Amides (Nylon)				X	X				
Cellulose Acetate			X	X	X		X		X
Cellulose Acetate Butyrate			X	X	X		X		X
Cellulose Nitrate			X				X	X	X
Ethyl Cellulose	X		X	X	X		X		X
Methyl Methacrylate	X		X	X	X		X		
Polvinyl Alcohol	X		X		X				
Polyethylene				X	X				X
Shellac			X					X	
Styrene Polymers			X	X	X		X		
Vinyl Aldehydes			X	X	X	X			
Vinylidene Chloride (Saran)			X	X	X				X
Vinyl Esters			X	X	X	X			

[1] Machines have been developed to permit the injection molding of thermosetting materials, but are not in common use at this time.

CHOICE OF A PLASTIC MATERIAL

A number of factors must be taken into account in order to arrive at the proper choice of a plastic material for a given application. These factors may be combined into four separate groups consisting of physical properties, electrical properties, eye appeal, and cost.

Physical properties, described in handbook form and in manufacturers' literature, have been carefully determined from laboratory samples tested according to the American Society for Testing Materials standards. Values will be found to cover a wide range for each material because of two principal factors:

1. The actual molding of a material under different conditions of temperatures and pressures will produce widely different results. For example, a tensile-strength range of from 5000 to 9000 pounds per square inch may be obtained with polystyrene. It is logical to assume that one should consistently obtain a strength figure of 7000 pounds per square inch in production.

2. Many materials are formulated to meet specific requirements requiring increased moldability, heat resistance, dimensional stability, etc. A handbook presentation of the characteristics for such a material must include all

of those values which can be obtained from each formulation. Cellulose acetate typifies this kind of a material, in that molding powders can be obtained having varying degrees of hardness, temperature resistance, and cold flow after molding.

Hence, variations may appear in the molding of test-specimen bars, from which the physical characteristics are determined. An additional variation appears in a molded item due to flow within the die during molding. It is apparent, therefore, that final judgment of a given material must be reserved for test results on the molded product, particularly if the safety factor is marginal.

Variation in the electrical properties comes about in much the same manner as those found to occur in the physical properties. However, these are of a less serious nature once the proper formulation is selected and the design is based upon its choice, whether such a choice involves dielectric constant, arc-tracking resistance, power factor, or dielectric strength.

The use of a plastic material, because of its beauty, involves an intangible factor in its choice which we have called "eye appeal." Eye appeal, at first sight, becomes sales appeal and results from styling, warmth or coolness of color, and "feel." The last factor can best be described as a finish which imparts life, is pleasant to touch, and is devoid of a dull or greasy luster. Once sales appeal has been effected, eye appeal must be supported by the physical characteristics of the material to maintain its surface finish under service conditions. This involves resistance to scratching, permanence of color, and dimensional stability.

The final determining factor for the use of any material is the cost of the finished product. Cost may be broken down into (1) capital investment, (2) personnel, (3) material cost, (4) hourly production, (5) finishing, and (6) packaging. If the choice lies between a suitable thermosetting material and a comparable thermoplastic purely on the basis of cost, one must pay particular attention to items (3), (4), and (5) above.

Actually, what is taking place today is an attempt to develop thermosetting-material equipment which will give the production obtainable from injection equipment used for thermoplastics and, in turn, thermoplastic-materials manufacturers are trying to approach the more important characteristics found in the thermosetting powders in the form of increased temperature resistance, higher impact values, and price reductions. Postwar production methods are certain to require analysis to be based solely on the particular item in question, and such an analysis must take into account all of those factors previously mentioned in order to obtain a clear-cut picture.

PROPERTIES OF MATERIALS

Tables VI, VII, and VIII indicate the outstanding properties of various materials; their physical properties, and their electrical properties, respectively. Table IX lists typical characteristics for laminated products. An average figure was chosen from handbook and manufacturers'-test data, and is tabulated as a single value for each characteristic. Since single values cannot indicate the possible range obtainable or available, these values must be used qualitatively.

TABLE VI

OUTSTANDING PROPERTIES FOR VARIOUS PLASTIC MATERIALS

Material	*Properties*
Allyl Resins	Clarity, hard surface, excellent optical properties. An impregnant for laminates.
Aniline Formaldehyde	Moisture and chemical resistance, improved dielectric properties.
Casein	Nonflammable, easy to machine and polish.
Lignin	Excellent extender in phenolics. Adhesive for laminates.
Melamine Formaldehyde	Hard surface, solvent resistance, arc-tracking resistance.
Phenol Formaldehyde	High impact, minimum cold flow, heat and chemical resistance.
Phenol Furfural	Initially developed as a substitute for phenol formaldehyde; has excellent molding qualities with heat, chemical, and moisture resistance.
Urea Formaldehyde	Flexural and tensile strength, solvent resistance, hardness and light-diffusion properties.
Amides (Nylon)	Heat resistance, chemical resistance, tough, moldable in very thin sections.
Cellulose Acetate	Good impact strength and mar resistance. Easy to mold.
Cellulose Acetate-Butyrate	Dimensional stability, smooth surface finish. High impact strength.
Cellulose Nitrate	Toughness, water resistance, easily fabricated.
Ethyl Cellulose	High impact strength, widely soluble.
Methyl Methacrylate	Cast sheets, good electrical properties, outdoor aging resistance, excellent optical properties.
Polydichlorostyrene	Characteristics similar to polystyrene plus high heat resistance.
Polyethylene	High impact, flexibility at low temperatures, excellent electrical properties.
Polystyrene	Excellent electrical properties, moisture resistance, optical properties.
Saran	High tensile strength. Chemical resistance.
Shellac	Scratch resistance, resilience, smooth finish.
Silicones	Resin used as impregnant with inorganic insulation. Provides high heat resistance, attendant with good electrical properties.
Styrene Copolymers	
Cerex	Physical properties equivalent to polystyrene, electrical properties better than most thermosetting materials, high temperature resistance.
Styraloy	Excellent low-temperature flexibility with heat and corona resistance.
Vinyls	Rigid forms to liquid resins. Adhesives, wire coatings, fabric coatings, sheets. Chemical resistance, flexibility at low temperature.

TABLE VII

TYPICAL PHYSICAL PROPERTIES FOR THERMOSETTING MATERIALS

Material	Impact-Izod; foot pounds per inch of notch	Compression strength; pounds per square inch	Tensile strength; pounds per square inch	Flexural strength; pounds per square inch	Maximum deflection; inches	Rockwell hardness;[1] *M*-scale	Continuous operating temperature; degrees Fahrenheit	Specific gravity	Approximate cost per pound
Allyl Resin									
Cast Stock	0.35	23,000	5500	9000	—	95	176	1.31	1.00
Aniline Formaldehyde	0.30	22,000	10,000	15,000	—	115	225	1.21	0.35
Casein	1.0	—	9000	—	—	—	160	1.35	—
Melamine Formaldehyde									
Alpha Cellulose	0.26	22,000	7500	14,000	0.062	120	210	1.49	0.52
Cotton Fabric	0.68	23,000	7000	14,000	0.049	110	225	1.40	0.55
Mineral Filled	0.31	20,000	6000	9300	0.021	110	350	1.98	0.53
Phenolics									
General Purpose	0.32	26,000	7000	11,000	0.060	113	250	1.38	0.18
Improved Impact	0.53	26,000	7500	12,000	0.065	100	250	1.36	0.26
Medium Impact	1.30	30,000	7500	12,000	0.068	100	240	1.38	0.31
High Impact	6.00	30,000	6500	13,000	0.095	100	230	1.35	0.43
Electrical	0.35	25,000	6000	9000	0.020	110	300	1.76	0.26
Chemical Resistant	0.20	27,000	6000	10,500	0.065	100	300	1.24	0.19
Heat Resistant	0.30	25,000	4200	7500	0.015	110	450	1.74	0.16
Urea Formaldehyde									
Alpha Cellulose	0.26	22,000	6000	15,000	0.065	120	180	1.45	0.35

[1] ASTM-D-48-39; ¼-inch-diameter ball, 10-kilogram static load, 100-kilogram added load. This value does not indicate surface-scratch resistance.

TYPICAL ELECTRICAL PROPERTIES FOR THERMOSETTING MATERIALS

Material	Volume resistivity; ohm-centimeter	Per cent power factor at 1 megacycle	Dielectric constant at 1 megacycle	Dielectric strength; volts per mil; ⅛-inch-thick sample; 60 cycles	Water absorption; per cent 24 hours
Allyl Resin					
Cast Stock	10^{11}	5.6	3.5	450	0.30
Aniline Formaldehyde	10^{12}	0.6	3.5	600	0.08
Casein	—	5.2	6.3	500	12.0
Melamine Formaldehyde					
Alpha Cellulose	10^{12}	2.9	7.0	300	0.7
Cotton Fabric	10^{11}	4.1	7.2	270	1.0
Mineral Filled	10^{13}	2.8	5.8	440	0.07
Phenolics					
General Purpose	10^{11}	4.0	5.0	400	0.7
Improved Impact	10^{10}	4.0	5.0	350	0.7
Medium Impact	10^{10}	6.0	6.0	300	1.0
High Impact	10^{9}	7.0	7.0	250	1.5
Electrical	10^{13}	0.85	4.0	450	0.03
Chemical Resistant	10^{10}	6.0	5.0	300	0.20
Heat Resistant	10^{10}	5.0	5.0	250	0.24
Urea Formaldehyde					
Alpha Cellulose	10^{10}	2.7	6.6	300	2.0

TABLE VIII

TYPICAL PHYSICAL PROPERTIES FOR THERMOPLASTIC MATERIALS

Material	Impact-Izod; foot-pounds per inch notch	Compression strength; pounds per square inch	Tensile strength; pounds per square inch	Flexural strength; pounds per square inch	Rockwell[1] hardness; *M*-scale	Continuous operating temperature; degrees Fahrenheit	Specific gravity	Approximate cost per pound
Amides (Nylon)	0.94	18,000	7000	12,500	90	250	1.14	$2.60
Cellulose Acetate	4.2	13,000	6000	10,000	40	160	1.3	0.48
Cellulose Acetate-Butyrate	5.0	15,000	6000	8000	30	170	1.2	0.53
Cellulose Nitrate	5.0	25,000	7500	9000	40	140	1.4	0.40
Ethyl Cellulose	6.5	11,000	7500	9500	40	170	1.1	0.70
Methyl Methacrylate	0.6	12,000	7000	15,000	90	155	1.2	0.85
Polyethylene	3.5	3000	1700	1700	29	200	0.92	0.85
Polyvinyl Alcohol	—	—	4000	—	—	200	1.25	—
Shellac	2.7	13,000	1500	—	—	175	1.90	0.15
Styrene Resins								
Polystyrene	0.4	14,000	7000	12,000	85	160	1.06	0.37
Polystyrene Base (Styramic)	0.3	11,000	3300	6500	72	180	1.36	0.70
Polydichlorostyrene (Styramic HT)	—	—	—	—	103	230	1.38	—
Styrene Copolymers								
Cerex	0.46	—	—	13,000	100	215	1.07	5.60
Styroloy	1.5	—	1000	—	—	145	0.96	—
Vinyls[2]	—	11,000	9000	12,000	20	130	1.30	0.75
Vinylidene Chloride	6.5	8000	5500	16,000	55	180	1.7	1.00

[1] This value does not indicate surface-scratch resistance.

[2] These materials are available in rigid to flexible forms, dependent upon compounding. Values given typify the flexible type of material.

TYPICAL ELECTRICAL PROPERTIES OF THERMOPLASTIC MATERIALS

Material	Volume resistivity; ohm-centimeter	Per cent power factor; 1 megacycle	Dielectric constant; 1 megacycle	Dielectric strength; volts per mil $\frac{1}{8}$-inch section; 60 cycles	Water absorption; per cent in 24 hours
Amides (Nylon)	10^{13}	5.5	4.0	400	1.5
Cellulose Acetate	10^{9}	3.0	4.0	375	4.0
Cellulose Acetate-Butyrate	10^{9}	3.0	4.7	325	1.7
Cellulose Nitrate	10^{11}	8.5	6.2	600	1.7
Ethyl Cellulose	10^{16}	2.5	3.2	600	1.5
Methyl Methacrylate	10^{15}	1.5	2.8	500	0.5
Polyethylene	10^{17}	0.03	2.3	900	0.01
Shellac	10^{9}	5.0	4.5	400	—
Styrene Resins					
Polystyrene	10^{18}	0.02	2.6	600	0.05
Polystyrene Base (Styramic)	—	0.04	2.5	—	0.046
Polydichlorostyrene (Styramic HT)	—	0.02	2.6	—	0.03
Styrene Copolymers					
Cerex	—	0.24	2.7	500	0.30
Styraloy	10^{20}	0.07	2.6	700	0.02
Vinylidene Chloride	10^{15}	4.0	4.0	350	0.09

TABLE IX

TYPICAL PHYSICAL PROPERTIES OF LAMINATES

Base material	Impact; foot-pounds per inch of notch	Compression strength; pounds per square inch	Tensile strength; pounds per square inch	Flexural strength; pounds per square inch	Rockwell hardness; *M*-scale	Water absorption; per cent 24 hours	Specific gravity
			Phenol Formaldehyde Resin				
Paper	1.0	26,000	11,000	18,000	100	2.5	1.34
Cotton Fabric	3.0	33,000	11,000	19,000	100	2.0	1.35
Asbestos Cloth	2.5	38,000	9000	18,000	95	1.5	1.60
Glass Fabric	6.5	44,000	17,000	23,000	110	0.40	1.50

TABLE IX—*continued*

TYPICAL PHYSICAL PROPERTIES OF LAMINATES

Base material	Impact; foot-pounds per inch of notch	Compression strength; pounds per square inch	Tensile strength; pounds per square inch	Flexural srrength; pounds per square inch	Rockwell hardness; *M*-scale	Water absorption; per cent 24 hours	Specific gravity
			Allyl Resin				
Paper	1.0	31,000	11,000	16,000	94	2.4	1.40
8-ounce Duck	3.6	29,000	7200	13,000	72	1.5	1.37
Glass Fabric	22.7	60,000	41,000	21,000	96	0.40	1.77
			Phenol Formaldehyde Bonded Wood Veneers				
1/45-inch Birch, phenolic film	12.4	20,000	37,000	6000	—	3.8	1.38
1/16-inch Maple, liquid resin	10.0	10,000	30,000	3700	—	8.8	1.37
$\frac{1}{8}$-inch Maple, impregnated	6.0	20,000	27,000	4000	—	4.0	1.30
Untreated Birch	—	6200	10,000	2000	—	—	0.63

TABLE X

TRADE NAMES AND MANUFACTURERS

	Allyl Resins
Allymer CR-39, CR-149, CR-170	Pittsburgh Plate Glass Company
MR-1, MR-1A, MR-17	Marco Chemicals, Inc.
	Amides
Nylon	E. I. Dupont Company
	Aniline Formaldehyde
Cibanite	Ciba Products Company
Dilectene	Continental Diamond Fiber Company
	Casein
Ameroid	American Plastics Corporation
Gala	George Morrell Corporation
Galorn	George Morrell Corporation
	Cellulose Acetate
Bakelite	Bakelite Corporation
Chemaco	Chemaco Corporation
Durashield	Plastics Fabricators, Inc.
Fibestos	Monsanto Chemical Company
Hercules	Hercules Powder Company
Lumapane	Celanese Celluloid Corporation
Lumarith	Celanese Celluloid Corporation
Nixonite	Nixon Nitration Works
Plastacele	E. I. DuPont Company
Tenite I	Tennessee Eastman Corporation
Textolite	General Electric Company
	Cellulose Acetate-Butyrate
Tenite II	Tennessee Eastman Corporation
Textolite	General Electric Company
	Cellulose Nitrate
Celluloid	Celanese Celluloid Corporation
Herculoid	Hercules Powder Company
Nitron	Monsanto Chemical Company
Nixonoid	Nixon Nitration Works
Pyralin	E. I. DuPont Company
Vimlite	Celanese Celluloid Corporation
	Ethyl Cellulose
Chemaco	Chemanaco Corporation
Ethocel	Dow Chemical Company
Ethofoil	Dow Chemical Company
Hercules E C	Hercules Powder Company
Lumarith E C	Celanese Celluloid Corporation
	Lignin
Benalite	Masonite Corporation
Catapak	Catalin Corporation
Lignolite	Marathon Chemical Company
	Melamine Formaldehyde
Catalin	Catalin Corporation
Melmac	American Cyanamid Company
Plaskon Melamine	Plaskon Company, Inc.
Resimene 803A	Monsanto Chemical Company
	Methyl Methacrylate (Acrylics)
Lucite	E. I. DuPont Company
Plexiglas	Rohm and Haas Company
Plexigum	Rohm and Haas Company
Textolite	General Electric Company
Vernonite (Denture)	Rohm and Haas Company
	Phenol Formaldehyde
Molding Compounds	
Bakelite	Bakelite Corporation
Co-ro-lite	Columbian Rope Company
Durez	Durez Plastics and Chemicals
Durite	Durite Plastics Company
Heresite	Heresite and Chemical Company
Indur	Reilly Tar and Chemical Corporation
Insurok	Richardson Company
Kys-ite	Keyes Fibre Company
Makalot	Makalot Corporation
Neillite	Watertown Manufacturing Company
Resinox	Monsanto Chemical Company
Textolite	General Electric Company
Resins	
Amberlite	Resinous Products and Chemical Company
Bakelite	Bakelite Corporation
Baker Resin	Baker Oil Tools, Inc.
Catabond	Catalin Corporation
Catalin	Catalin Corporation
Catavar	Catalin Corporation
Gemstone	A. Knoedler Company
Haveg	Haveg Corporation
Indur Varnish	Reilly Tar and Chemical Corporation
Marblette	Marblette Corporation
Opalon	Monsanto Chemicals Company
Prystal	Catalin Corporation
Tego	Resinous Products and Chemicals Company
	Phenol Furfural
Durite	Durite Plastics
Resin X	Plastics Industry Technical Institute
	Polyethylene
Polythene	E. I. DuPont Company
Polyethylene	Carbide and Carbon Chemical Company
	Polystyrene and Derivatives
Amphenol	American Phenolic Corporation
Bakelite	Bakelite Corporation
Cerex	Monsanto Chemicals Company
Chemaco	Chemaco Corporation

TABLE X—*continued*

TRADE NAMES AND MANUFACTURERS

	Polystyrene and Derivatives—continued
Loalin	Catalin Corporation
Lustron	Monsanto Chemical Company
Plax	Plax Corporation
Styramic X214	Monsanto Chemical Company
Styramic HT	Monsanto Chemical Company
Styraloy	Dow Chemical Company
Styrofoam	Dow Chemical Company
Styron	Dow Chemical Company
	Polyvinyl Acetal
Alvar	Shawinigan Products Corporation
	Polyvinyl Acetate
DuPont Polyvinyl	E. I. DuPont Company
Gelva	Shawinigan Products Corporation
Vinylite A	Carbide and Carbon Chemical Company
Vinylseal	Carbide and Carbon Chemical Company
	Polyvinyl Alcohol
Compar	Resistoflex Corporation
PVA	E. I. DuPont Company
Resistoflex	Resistoflex Corporation
	Polyvinyl Butyral
Buticite	E. I. DuPont Company
Butvar	Shawinigan Products Corporation
Saflex	Monsanto Chemical Company
Vinylite X	Carbide and Carbon Chemical Company
	Polyvinyl Chloride
Geon 100 Series (Koroseal, etc.)	B. F. Goodrich Company
Vinylite Q	Carbide and Carbon Chemical Company
	Polyvinyl Chloride Acetate
Vinylite V	Carbide and Carbon Chemical Company
Vinyon	Carbide and Carbon Chemical Company
	Polyvinyl Formal
Formex	General Electric Company
Formvar	Shawinigan Products Corporation
	Shellac
Complac	Poinsettia, Incorporated
Compo-Site	Compo-Site, Incorporated
Harvite	Siemon Company
Lacanite	Consolidated Molded Products Company
	Silicones
Silicones	Dow-Corning Company
	Urea Formaldehyde
Bakelite	Bakelite Corporation
Beetle	American Cyanamid Company
Insurok	Richardson Company
Makolot	Makolot Corporation
Plaskon	Libby-Owens-Ford Glass Company
Textolite	General Electric Company
Uformite	Resinous Products and Chemicals Company
Urac	American Cyanamid Company
	Vinyls—General
Chemaco	Chemaco Company
Gemflex	Gemloid Corporation
Gemloid	Gemloid Corporation
Synflex	Industrial Synthetics Corporation
Vynate	Monsanto Chemical Company
	Vinyl Vinylidene Chloride
Geon 200 Series	B. F. Goodrich Company

TABLE X—*continued*

TRADE NAMES AND MANUFACTURERS

	Vinylidene Chloride
Saran	Dow Chemical Company
Vec	Pierce Plastics, Incorporated
Velon	Firestone Rubber and Laytex Company
	Laminates
Aqualite	National Vulcanized Fibre Company
Celoron	Continental-Diamond Fibre Company
Dilecto	Continental-Diamond Fibre Company
Duraloy	Detroit Paper Products Corporation
Farlite	Farley and Loetscher Manufacturing Company
Formica	Formica Insulation Company
Insurok	Richardson Company
Lamicoid	Mica Insulator Company
Lamitex	Franklin Fibre-Lamitex Corporation
Micarta	Westinghouse Manufacturing Corporation
Ohmoid	Wilmington Fibre Specialty Company
Panelyte	St. Regis Paper Company
Papreg	Forest Products Company
Phenol Fibre	Penn Fibre and Specialty Company
Phenolite	National Vulcanized Fibre Company
Plastiply	Haskelite Manufacturing Corporation
Pluswood	Pluswood, Incorporated
Plymetl	Haskelite Manufacturing Corporation
Plymold	Haskelite Manufacturing Corporation
Ply-Tech	Technical Ply-Woods Company
Polyflex	Plax Corporation
Preg-Tech	Technical Ply-Woods Company
Pregwood	Formica Insulation Company
Resnprest	Plylock Corporation
Revolite	Atlas Powder Company
Spauldite	Spaulding Fibre Company
Synthane	Synthane Corporation
Taylor	Taylor Fibre Company
Textolite	General Electric Company
Ucinite	United-Carr Fastener Company

TABLE XI

TRADE NAMES AND COMPOSITION

Allymer CR-39, CR-149, CR-170	Allyl Resin
Alvar	Polyvinyl Acetal
Amberlite	Phenol Formaldehyde
Ameroid	Casein
Amphenol	Polystyrene
Aqualite	Phenolic Laminates
Bakelite	Thermosetting and Thermoplastic Materials
Baker Resin	Phenol Formaldehyde
Beetle	Urea Formaldehyde
Benalite	Lignin Sheet
Butacite	Polyvinyl Butyral
Butvar	Polyvinyl Butyral
Catabond	Phenol Formaldehyde
Catalin	Phenol Formaldehyde, Urea Formaldehyde
Catapak	Lignin
Catavar	Phenol Formaldehyde Varnish
Celluloid	Cellulose Nitrate
Celoron	Thermosetting Laminates
Cerex	Polystrene Copolymer
Chemaco	Thermoplastic Materials
Cibanite	Aniline Formaldehyde
Compar	Polyvinyl Alcohol Base
Complac	Shellac
Compo-site	Shellac
Co-ro-lite	Phenol Formaldehyde
Dilectene	Aniline Formaldehyde
Dilecto	Phenolic Laminates
DuPont Polyvinyl	Polyvinyl Acetate
Duraloy	Phenolic Laminates

TABLE XI—*continued*
TRADE NAMES AND COMPOSITION

Durashield	Cellulose Acetate
Durez	Phenolic Laminates
Durite	Phenol Formaldehyde, Phenol Furfural
Ethocel	Ethyl Cellulose
Ethofoil	Ethyl Cellulose
Farlite	Phenolic Laminates
Fibestos	Cellulose Acetate
Formex	Polyvinyl Formal
Formica	Phenolic Laminates
Formvar	Polyvinyl Formal
Gala	Casein
Galorn	Casein
Gelva	Polyvinyl Acetate
Gemflex	Vinyls
Gemloid	Vinyls
Gemstone	Phenol Formaldehyde
Geon 100 Series	Polyvinyl Chloride
Geon 200 Series	Vinyl Vinylidene Chloride
Haveg	Phenol Formaldehyde
Harvite	Shellac
Hercules	Cellulose Acetate, Ethyl Cellulose
Herculoid	Cellulose Nitrate
Heresite	Phenol Formaldehyde
Indur	Phenol Formaldehyde
Insurok	Phenolic and Urea Laminates
Koroseal (Geon 100 Series)	Polyvinyl Chloride
Korogel (Geon 100 Series)	Polyvinyl Chloride
Kys-ite	Phenol Formaldehyde
Lacanite	Shellac
Lamicoid	Phenolic Laminates
Lamitex	Phenolic Laminates
Lignolite	Lignin
Loalin	Polystyrene
Lucite	Methyl Methacrylate
Lumapane	Cellulose Acetate
Lumarith	Cellulose Acetate, Ethyl Cellulose
Lustron	Polystyrene
Makalot	Phenol Formaldehyde, Urea Formaldehyde
Marblette	Phenol Formaldehyde
Melmac	Melamine Formaldehyde
Melurac	Melamine-Urea Resin
Micarta	Phenolic Laminates
MR1, MR1A, MR-17	Allyn Resin
Neillite	Phenol Formaldehyde
Nitron	Cellulose Nitrate
Nixonoid	Cellulose Nitrate
Nixonite	Cellulose Acetate
Nylon	Polyamide
Ohmoid	Phenolic Laminates
Opalon	Phenol Formaldehyde
Panelyte	Phenolic Laminates
Papreg	Laminated Paper
Phenol Fibre	Phenolic Laminates
Phenolite	Phenolic Laminates
Plaskon	Urea Formaldehyde, Melamine Formaldehyde
Plastacele	Cellulose Acetate
Plastiply	Resin Covered Plywood
Plexiglas	Methyl Methacrylate
Plexigum	Methyl Methacrylate
Plax	Polystyrene
Pluswood	Impregnated Plywood
Plymetl	Metal Bonded to Plywood
Plymold	Resin Bonded Plywood
Ply-Tech	Resin Bonded Plywood
Polyflex	Thermosetting Laminates
Polythene	Polyethylene
Preg Tech	Impregnated Wood
Pregwood	Impregnated Wood
Prystal	Impregnated Wood
PVA	Polyvinyl Alcohol
Pyralin	Cellulose Nitrate
Resimene 803A	Melamine Formaldehyde
Resin X	Liquid Furfural Resin
Resinox	Phenal Formaldehyde
Resistoflex	Polyvinyl Alcohol
Resnprest	Resin Bonded Plywood
Revolite	Phenolic Laminates
Saflex	Polyvinyl Butyral
Saran	Vinylidene Chloride
Silicons	Organo-Silicon Oxide Polymers
Spauldite	Phenolic Laminates
Styramic HT	Polydichlorostyrene
Styrofoam	Expanded Polystyrene
Styron	Polystyrene
Styraloy	Elastomeric Polystyrene Copolymer
Styramic X214	Polystyrene Base
Synflex	Vinyls
Synthane	Phenolic Laminates
Taylor	Phenolic Laminates
Tego	Phenolic Impregnated Tape
Tenite I	Cellulose Acetate
Tenite II	Cellulose Acetate—Butyrate
Textolite	Thermosetting and Thermoplastic Materials
Ucinite	Phenolic Laminates
Urformite	Urea Formaldehyde
Urac	Urea Formaldehyde
Vec	Vinylidene Chloride Resins
Velon	Vinylidene Chloride Resins
Vernonite	Methyl Metacrylate (Denture)
Vimlite	Cellulose Nitrate
Vinylite	Vinyl Materials
Vinylseal	Polyvinyl Acetate
Vinyon	Polyvinyl Chloride Acetate
Vynate	Vinyl Materials

CONCLUSION[1]

No attempt has been made to present a comprehensive review of the plastics field, but, rather, to offer a progressive series of tabulations for the materials in common use today and to show their classifications, one with respect to another. From these general characteristics one can choose those materials which appear to offer possibilities for the application under consideration and obtain their specific characteristics from manufacturers' test information.

In conclusion, plastics should be used only on the basis of their own merits; utility should not be sacrificed for cost. Observance of such a policy will establish a firm footing for their continued use and the public will not come to regard plastics as a group of low-cost substitutes.

APPENDIX

EXPANSION COEFFICIENTS

The use of metal inserts in molded plastics is common practice, wherein the insert is set into position in the die prior to actual molding. The efficiency of such an insert depends upon its function and the type of requirements placed upon the bond between the metal and the plastic.

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[1] The trade names or trade-marks and the identity of the products and manufacturers listed herein were developed from the best sources available to the writer. No responsibility is assumed for any error in spelling or in the identity of the product or source, or for any omissions of individual products or manufacturers.

When the bond is one of a mechanical nature, design is confined to the type and amount of serration or perforation to be used on the metal insert in keeping with the flow characteristics of the plastic material. However, if the bond must also withstand liquid or gas pressures, the difference in expansion coefficients existing between the metal and the plastic must be given serious consideration. From the list of coefficients for common materials, given in Table XII, it can be seen that definite limitations exist, and generally, some kind of a third material must be used between the plastic and metal to serve as a compliant, gas tight member.

TABLE XII

EXPANSION COEFFICIENTS PER 100 DEGREES CENTIGRADE

Alluminum	0.0023	Ethyl Cellulose	0.012
Copper	0.0017	Melamine Formaldehyde	0.002
Glass	0.00085	Methyl Methacrylate	0.008
Mercury	0.018	Nylon	0.010
Porcelain	0.00036	Phenol Formaldehyde	0.005
Silver	0.0019	Polyethylene	0.019
Steel	0.0012	Polystyrene	0.007
Cellulose Nitrate	0.011	Saran	0.016
Cellulose Acetate	0.012	Urea Formaldehyde	0.003

LIGHT TRANSMISSION AND OPTICS

Certain members of the thermosetting group have become well recognized for their light-diffusing properties. For example, urea formaldehyde has been used to make lamp reflector-bowls, and marketed in large quantities.

Acetate film lends itself well to the making of color screens, due to the uniformity of color and the ease with which varying shades can be controlled.

Methyl methacrylate and polystyrene have the required clarity to transmit light around curves, which property is utilized in medical instruments, edge-lighted signs, instrument panels, etc.

Edge-lighted dials obtain their effect by the use of offset numerals and designs, which pick up the light to present a controlled illuminated area. Such parts may be compression-molded to avoid stress and flow lines which might be apparent in an injection-molded piece.

Polystyrene is injection-molded into faceted forms for costume jewelry. It is superior to methacrylate for such items because of its greater scratch resistance and surfact hardness.

The most critical of all applications involves the manufacture of optical lenses. Injection-molding is satisfactory for the manufacture of common magnifying glasses; whereas, compression-molding techniques must be used for higher-quality lenses because of the greater freedom from internal stress. Multiple-element lens systems which must meet rigid optical specifications are made either by the grinding of the lens from cast blanks or the direct casting of the lens, using optically ground molds. The most common materials for these applications are the methacrylates and the polystyrenes. Allyl resins offer excellent possibilities for the casting of lenses when the material becomes available, since it provides greater abrasion resistance.

The principal precautions that must be observed in the manufacture and the use of plastic lenses involve (1) stress relief and complete polymerization; (2) parts must be mounted so as to allow for expansion and contraction with temperature; (3) housings for the lens assemblies should be dust-proof; and (4) extreme care must be exercised if cleaning of a particular surface is necessary.

The use of the methacrylates or polystyrene for reflecting mirrors falls into the same category of equipment as lenses. Such pieces can be readily molded or cast, using an optical surface on the male part of the die. The current problem involves deposition of the reflecting metal film on the plastic. Once this is accomplished economically, quantity-production reflector costs can be reduced below the cost for large, all-metal, plated reflectors or silvered-glass mirrors in use today.

LAMINATES

Flat sheets, rods, and tubes built up from paper, fabric, and wood veneers, using various animal and vegetable glues, have been available for some years. However, during the past several years, new synthetic-resin binders, new base materials, and new methods of fabrication have resulted in vastly improved products.

Generally speaking, the fabrication of laminated products makes use of one of the following methods: (1) high pressure (1500 to 3000 pounds per square inch), 300 degrees Fahrenheit, steam heat, press cure; (2) high pressure, radio-frequency heat curve; (3) low pressure (15 to 75 pounds per square inch), bag molding, steam autoclave-cure, 300 degrees Fahrenheit; (4) contact pressure (15 pounds per square inch or less), 140 to 250 degrees Fahrenheit oven cure.

In the use of the high-pressure method, materials may be resin bonded and/or resin impregnated. Wood veneers 1/80- to 1/30-inch thick are bonded by resin film placed between the layers of ply; veneers 1/30- to 1/16-inch in thickness are usually coated with a liquid resin; whereas, thicknesses of 1/16- to 1/8-inch must be impregnated by complete immersion. The technique is, therefore, dictated by the amount of resin that can be forced into a given thickness of material.

After stacking the veneers between the press platens, thermocouples are placed between the plies throughout the stack height, to permit temperature control on the outer surfaces and to determine when the proper temperature is reached at the center of the stack. When the resin has cured, the press temperature must be reduced so as to permit the center of the stack to reach approximately 140 degrees Fahrenheit before the pressure can be released. Curing time approximates one hour per inch of thickness. Radio-frequency heating must be used for sections greater than two inches thick.

Low-pressure and contact-pressure molding have been made feasible by the development of resins which have higher penetrating power and will set up with or without the application of external heat. The reduction of time, temperature, and pressure has resulted in the economical production of large, laminated, curved structures approaching the strengths of sheet metals. Complete boat hulls and plane sections made in this manner have given excellent service in battle engagements.

In the bag-molding method, any of the resin-impregnated materials previously indicated in Table III may be formed around a male core or within a female die. The die may be made from wood, metal, cast resin, or plaster, depending upon expected service life. A rubber bag or "blanket" is then used to act as the mating member of the die, either as the female or the male part, respectively. Upon evacuation of the air between the "blanket" and the material being formed, the bag serves to transmit a uniform pressure over the surface of the material to be formed and cured, and, furthermore, prevents contact of the resin with live steam in the autoclave, when used. Other methods of holding the wrapped material in place without the use of the bag process have been developed.

Low-pressure and contact molding have been developed during the past several years, and their use has been confined to war applications. They offer possibilities in the furniture field sufficient to cause considerable concern among the furniture manufacturers at this time, and it is certain that future applications depend only upon the ingenuity of the molder.

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Effect of Surface Finish and Wall Thickness on the Operating Temperature of Graphite Radio-Tube Anodes*

L. L. WINTER†, MEMBER, I.R.E., AND H. G. MACPHERSON†

Summary—The effect of rough and smooth surfaces and variable wall thicknesses on the operating temperature of radio-tube graphite anodes has been determined. Simultaneous measurements of internal and external temperatures were made with two optical pyrometers on rough and polished hollow graphite cylinders with variable wall thicknesses. Because of the ideal black-body conditions inside the graphite tubes, true internal temperatures were obtained. The optical pyrometer sighted on the outside of the cylinder measured the radiation at a wave length of 6530 angstrom units in terms of brightness temperature. From these data, calculations were made showing the true temperature differences and differences in emissivity and in radiation of the various graphite cylinders.

Observations made show that varying the wall thickness of these graphite cylinders from 0.07 to 0.10 inch gave a difference of not more than 20 degrees centigrade in operating temperature. Polishing the surface of the graphite reduced the emissivity by 8 per cent, and increased the surface temperature by 25 degrees centigrade over that obtained for the rough-surface cylinder at 1000 degrees centigrade. Radiation from these surfaces indicates that a roughened anode surface at 1000 degrees centigrade has a radiation capacity increased over that of the polished anode by 1.1 watts per square centimeter.

GRAPHITE anodes for some recent radio-tube applications require a smooth surface finish in order to eliminate surface irregularities and remove loosely held particles.

In this connection, it became desirable to measure and to evaluate the effects of smooth and rough anode finishes and of varying anode wall thicknesses on the operating temperatures. Because of the favorable radiation characteristics of graphite, and since few anodes operate at temperatures higher than 1000 degrees centigrade, measurements were made between 850 and 1075 degrees centigrade, a range in which optical measurements are quite reliable. The general procedure followed was to heat a graphite cylinder in a vacuum to incandescence by passing current through it and make simultaneous measurements of both internal and external brightness temperatures at an effective wave length of 6530 angstrom units.

Graphite cylinders 12 inches long by 1-inch outside diameter were polished to give a shiny surface, and

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similar-size cylinders were sanded to give as dull a surface as possible. Cylinders having the desired surface characteristic were machined to wall thicknesses of 0.100 and 0.070 inch. The photograph in Fig. 1 shows the difference in surface smoothness and in wall thickness of the cylinders.

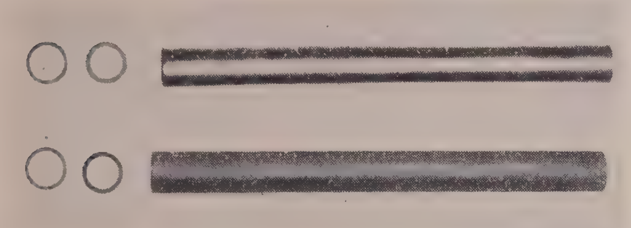

Fig. 1—Comparison of graphite cylinders having smooth and rough surfaces and thick and thin walls.

The graphite cylinders were mounted axially in the pyrex chamber shown in Fig. 2, which was evacuated to a pressure of less than 5 microns to eliminate any oxidation of the graphite and reduce convection heat losses, The other details of the arrangement are indicated in the diagram. The power input was adjusted so that the cylinder came up to operating temperature in 80 to 100 seconds.

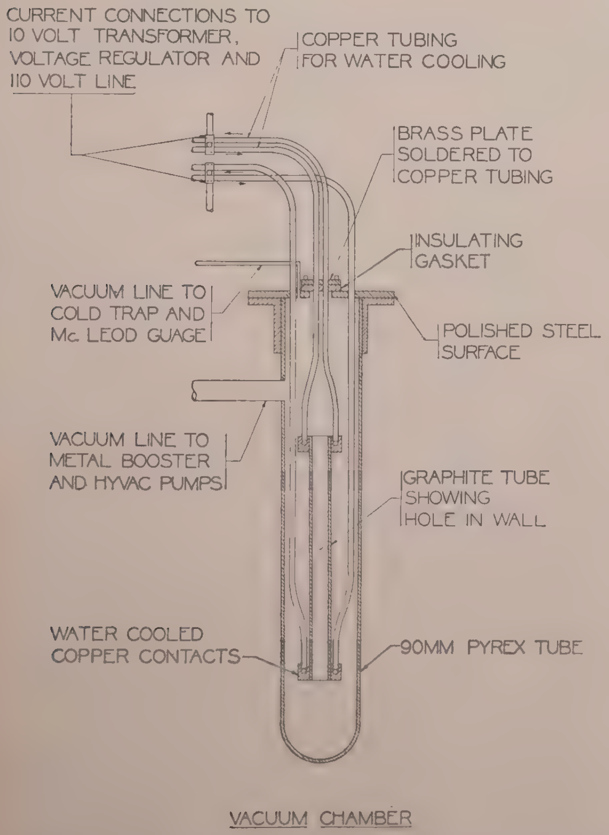
CURRENT CONNECTIONS TO 10 VOLT TRANSFORMER, VOLTAGE REGULATOR AND 110 VOLT LINE
COPPER TUBING FOR WATER COOLING
BRASS PLATE SOLDERED TO COPPER TUBING
INSULATING GASKET
VACUUM LINE TO COLD TRAP AND Mc LEOD GUAGE
POLISHED STEEL SURFACE
VACUUM LINE TO METAL BOOSTER AND HYVAC PUMPS
GRAPHITE TUBE SHOWING HOLE IN WALL
WATER COOLED COPPER CONTACTS
90MM PYREX TUBE
VACUUM CHAMBER

Fig. 2—Vacuum-chamber assembly showing the graphite cylinder held in position inside the pyrex tubes by means of water-cooled current conductors.

The temperature inside of the graphite cylinder was measured by an optical pyrometer sighted through a 0.16-inch-diameter hole drilled in the cylinder wall. A simultaneous measurement was made of the brightness temperature of the outside of the cylinder with a pyrometer sighted on the outside wall opposite the hole. The radiation coming through the hole was very nearly black-body radiation, since the cylinder is heated to a uniform temperature for about 6 inches of its length. The optical pyrometers were calibrated over the temperature range 750 to 1075 degrees centigrade against a standard chromel-alumel thermocouple.

The results of the temperature measurements on the graphite cylinders are plotted in Fig. 3, where the difference between the inside and outside temperatures is indicated as a function of the inside temperature of the cylinder. The optical pyrometers were interchanged once during each set of measurements to insure that the temperature differences observed would be independent of any differences in the calibrations of the two instruments. The circles represent temperatures obtained with one arrangement of the optical pyrometers, and the crosses those determined with the other arrangement. It is apparent that there is a systematic difference between the readings with the two pyrometers,

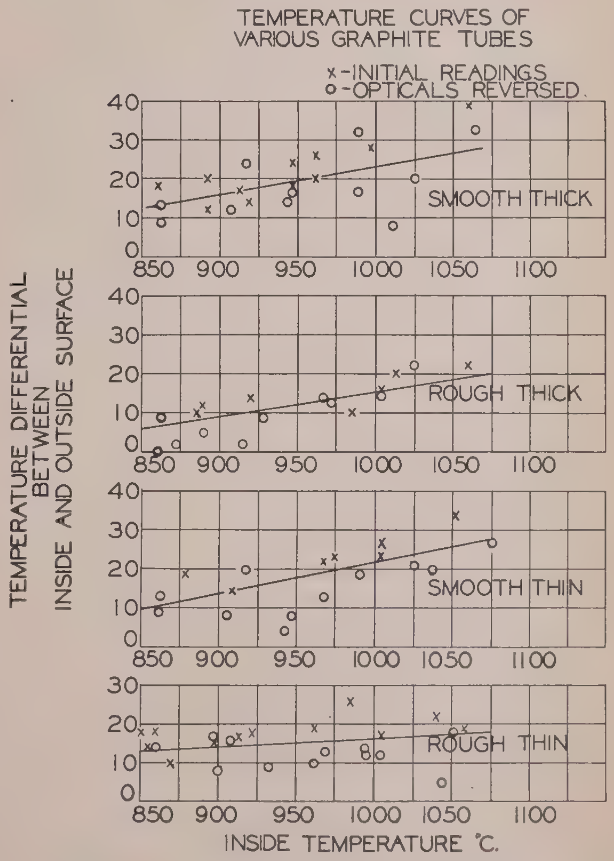

Fig. 3—These curves indicate the temperature differential between the inside and outside surface of the various graphite cylinders over an inside temperature range of 850 to 1075 degrees centigrade.

due to a slight error in calibration. However, this does not enter into the final results, because the method of interchanging the pyrometers eliminates this error. When plotted on this expanded scale, the temperature measurements show a considerable spread. However, the probable error of any one temperature-difference measurement is estimated to be not more than 5 degrees centigrade, so that when a number of points are averaged, the accuracy is sufficient to permit pointing out the differences between the cylinders. In general, the difference in temperature between the inside and outside surface increases with the temperature level but the maximum difference, even at 1050 degrees centigrade with the thickest of the cylinders, is only about 30 degrees centigrade.

Table I gives the average temperature differences obtained from the various cylinders and selected regions of temperature. These average values are probably the most objective means of judging the differences between the samples, since they represent fairly large groups of measurements. In obtaining the averages, the points for each pyrometer arrangement were averaged separately and the mean of the two averages was taken

TABLE I

	Average Temperature Difference	
	In Range 850°–950° C.	In Range 950°–1050° C.
Rough dull surface		
Thin wall (0.066 inch)	14.0° C.	16.0° C.
Thick wall (0.102 inch)	8.4° C.	14.6° C.
Smooth shiny surface		
Thin wall (0.070 inch)	13.4° C.	21.1° C.
Thick wall (0.102 inch)	16.1° C.	22.6° C.

The effect of the thickness of the cylinder walls can be evaluated readily. In the 850-to-950-degree temperature region, the average difference of temperature for the thin-wall cylinders was 13.7 degrees, and for the thick-wall cylinders, 12.3 degrees. In the 950-to-1050-degree range, the value for the thin wall was 18.5 degrees and for the thick wall, 18.6 degrees. The indicated difference between the two wall thicknesses is less than the error of measurement, and in one case is in the reverse direction from that expected, so it is obvious that there is no appreciable difference in the temperature drop for the two wall thicknesses.

Calculations of the temperature differences based on heat conduction check this observation. In our experiment, the temperature difference between the outside and the inside of the cylinder is approximately one half of the value which would obtain in a radio-tube anode. This is due to the difference in the mode of heating employed. With our method of heating by the I^2R loss in the cylinder, the heat is generated uniformly throughout the entire graphite wall and is, on the average, conducted only halfway through the cylinder wall. In the operation of a radio tube, on the other hand, the heat is introduced at the inside surface and must all be conducted through the entire thickness of the graphite.

The temperature differences between the inside and outside surfaces were calculated for both cases, (1) that of uniform heating and (2) that of heating at the inside surface, from the following formulas:

$$\Delta T = (Qr_2/k)(1/2 - r_1^2/(r_2^2 - r_1^2) \ln r_2/r_1) \qquad (1)$$

$$\Delta T = (Qr_2/k) \ln r_2/r_1. \qquad (2)$$

In these formulas

ΔT = the true temperature difference between the inside and outside surface of the cylinder.

Q = the heat to be dissipated per unit area of the outside surface of the cylinder.

k = the thermal conductivity of graphite at the operating temperature of the cylinder.

r_1 and r_2 = the inner and outer radii of the cylinder.

Table II gives the calculated temperature differences between the inner and outer surfaces at various operating temperatures of the cylinder. In the calculation, it was assumed that the heat dissipated from the outside surface of the cylinder was equal to the radiation of a body with an emissivity of 0.9 at its operating temperature. The values of thermal conductivity used were 0.080, 0.071, and 0.062 gram calories per second per centimeter squared per degree centigrade per centimeter, for the three temperatures, 800 degrees centigrade, 900 degrees centigrade, and 1000 degrees centigrade, respectively. These values are estimated as probable values for the transverse thermal conductivity of the graphite used. The conductivity at 800 degrees centigrade is approximately 50 per cent of the longitudinal value given by Powell.[1]

TABLE II

Calculated Temperature Difference between Inner and Outer Surfaces

	Operating Temperature of Cylinder		
	800° C.	900° C.	1000° C.
A. Heat generated within the walls by resistance heating			
Thin wall (0.070 inch)	1.8° C.	3.0° C.	4.7° C.
Thick wall (0.100 inch)	2.8° C.	4.4° C.	7.0° C.
B. Heat generated at inside surface			
Thin wall (0.070 inch)	3.9° C.	6.1° C.	9.7° C.
Thick wall (0.100 inch)	5.7° C.	9.1° C.	14.5° C.

From Table II it can be seen that, in this experiment, the maximum difference to be expected between the thin- and thick-wall cylinders is 2.3 degrees at 1000 degrees centigrade. This calculation is therefore in agreement with our observation that the temperature difference is small. Even though the temperature difference in the case of a radio-tube anode would be approximately twice this, it is still too small to be important.

Examination of Table I shows that there is a real difference between the operation of the bright- and dull-finish cylinders. This difference can best be evaluated by calculating an emissivity for the two surfaces. First, however, we must subtract the real temperature difference between the inside and outside of the tube walls, due to the conduction of heat from the inside to the outside, and then the remaining temperature difference will be only an apparent difference, due to the fact that the emissivity of the surface is less than 1.0. We have used the calculated values of the true temperature difference given in Table II for making this subtraction because they are more accurate than any values that could be calculated from the data. Table III gives the values of the apparent temperature difference after the real temperature difference has been subtracted. Values for the thin- and thick-wall cylinders are averaged together since the effect of the wall thickness has been eliminated by the subtraction.

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[1] R. W. Powell, "The thermal and electrical conductivities of a sample of acheson graphite from 0° to 800° C.," *Proc. Phys. Soc.* (London), vol. 49, pp. 419–426; 1937.

TABLE III

Apparent- or Brightness-Temperature Difference between the Inside and Outside of Cylinders Due to Emissivity Value

	In Range 850°–950° C.	*Average*	In Range 950°–1050° C.	*Average*
Rough dull surface				
Thin wall	11.0	7.5	11.3	9.5
Thick wall	4.0		7.6	
Smooth shiny surface				
Thin wall	10.4	11.0	16.4	16.0
Thick wall	11.7		15.6	

For each value of this apparent-temperature difference, an emissivity was calculated from the following equation:

$$\log_{10} e = 0.4343 c_2 \Delta T / \lambda T^2.$$

In this equation

e = the emissivity

c_2 = the second radiation constant which has a value of 1.432 centimeter—degree centigrade

λ = the wave length in centimeters of the radiation used for the brightness measurement

ΔT = the apparent difference in temperature

T = the operating temperature of the cylinder.

Values of the emissivity calculated in this way are given in Table IV.

TABLE IV

Calculated Emissivities

	Data at 900° C.	Data at 1000° C.	Average
Rough dull surface	0.90	0.88	0.89
Smooth shiny surface	0.84	0.805	0.82

The average calculated value of the emissivity for the sanded-graphite surface is about 0.89, and for the shiny surface about 0.82. Although these emissivities are not claimed to be very accurate, the values are quite reasonable in view of previous experience with graphite. The emissivities calculated in this way are spectral emissivities for radiation of effective wavelength 0.653 microns. However, there is every reason to believe that carbon is a true grey body having the same emissivity at all wave lengths and temperature, so that spectral and total emissivities can be used interchangeably.[2]

The difference in emissivity of the two surfaces can be used to calculate the difference in operating temperature that would result from the use of the two types of surface. By the Stefan-Boltzmann radiation law, the energy radiated is proportional to the product of the emissivity and the fourth power of the absolute temperature. For example, the roughened, dull-finish-graphite surface will radiate 6.7 watts per square centimeter of surface if it operates at a temperature of 800 degrees centigrade. To radiate the same heat energy, the smooth shiny-surface anodes would operate at a temperature of 822 degrees centigrade, or 22 degrees higher. Similarly, to radiate the same amount of heat as the rough-surface anode at 1000 degrees centigrade, the smooth shiny anode would have to operate at a temperature of 1025 degrees centigrade.

Table V gives the radiation from rough- and polished-graphite cylinder surfaces at 900 degrees and 1000 degrees centigrade.

TABLE V

Surface condition	Emissivity	Temperature	Radiation watts per square centimeter
Smooth polished surface	0.82	900° C.	8.8
Rough dull surface	0.89	900° C.	9.6
Smooth polished surface	0.82	1000° C.	12.2
Rough dull surface	0.89	1000° C.	13.3

From these experiments, it is evident that differences of not more than 30 degrees centigrade will be found in radio-tube anodes operating at about 1000 degrees centigrade as a result of differences in type of surface and wall thicknesses of 0.070 to 0.100 inch.

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[2] American Institute of Physics, "Temperature, Its Measurement and Control in Science and Industry," Reinhold Publishing Corp., New York, N. Y., 1941; comment by H. T. Wensel, p. 1149.

A Coil-Neutralized Vacuum-Tube Amplifier at Very-High Frequencies*

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Summary—This paper describes a two-stage single-side coil-neutralized amplifier employing an experimental triode operating in the vicinity of 140 megacycles. Circuit features are described and typical operating conditions are indicated. Typical distortion characteristics at low power levels are also included.

PART I. INTRODUCTORY CONSIDERATION OF THE PROPERTIES OF THIS TYPE OF NEUTRALIZATION

WELL KNOWN are the usual methods employed in the neutralization of vacuum-tube amplifiers such as the "capacitance bridge," the Rice, and the Hazeltine arrangements. These methods use the general principle of neutralizing the voltage fed back from the output circuit to the input circuit through an inherent branch circuit by the addition of an auxiliary branch circuit which returns to the same input point an equal voltage of opposite phase.

A less familiar method using a variation of his principle was proposed by H. W. Nichols[1,2] nearly two decades ago. In this method the grid-to-plate capacitance of a vacuum tube was antiresonated with a suitable inductance at the operating frequency, thereby greatly reducing the admittance coupling the output and the input circuits. For convenience, we shall refer to this method as "coil neutralization." Because of a steadily increasing demand for power-amplifier stages in the ultra-high-frequency range, for which no adequately neutralized power tube was available,[3] and because of the greater technical difficulties encountered in applying conventional neutralizing schemes at ultra-high frequencies, an experimental study was begun to determine the operating characteristics of coil-neutralized triodes at very-high frequencies.

A qualitative consideration of the advantages and disadvantages of this method of neutralization may serve to orient the reader who is not familiar with the method. Coil neutralization has the disadvantage that careful adjustments must be made for each operating frequency. These adjustments, when made, are, however, reliable and duplicable at will. Minor readjustments are required only occasionally to maintain optimum neutralization, and this is true even for tube replacements. This inherent drawback loses significance in fixed-frequency services for which the ultra-high frequencies are particularly applicable. One important advantage of this method is the increased simplicity in circuit arrangement which may be achieved for either single-side or push-pull applications. The use of triodes with grid and plate terminals adjacent to each other greatly simplifies the mechanical layout of the circuit elements.

It is appreciated in a qualitative way that the capacitance to be antiresonated between the grid and plate elements includes not only the direct interelectrode capacitance, but also the total stray capacitance between these elements as well. Furthermore, the inherent inductance in plate and grid leads merely contributes to the total inductance required to antiresonate the total grid-to-plate capacitance. In this method, therefore, two factors which disturb conventional neutralizing technique have been put to service. At very-high frequencies the inductances and capacitances inadvertently introduced into a neutralizing circuit by an element such as a capacitor often result in unsatisfactory or just tolerable neutralization, and may also contribute to the establishment of parasitic oscillations. The ultimate high-frequency application of coil neutralization takes the form of a short transmission line, either of the open $1/2\lambda$ type, or the closed $1/4\lambda$ type, between the grid and plate elements. No blocking capacitor is required if the $1/2\lambda$ line is used. In general, in the very-high- and ultra-high-frequency region, less stray capacitance will be added to the circuit with coil neutralization than with conventional neutralizing means. Such direct benefits as less capacitance current, a broader transmission band, and a higher frequency of operation result.

When coil neutralization is applied to a tuned-grid, tuned-plate triode, a sufficient condition for nonoscillation is that at any frequency the impedances of the three circuits must be of like nature, resistive, inductive, or capacitive. At the particular frequency for which the coil and the grid-to-plate capacitance antiresonate, the impedance between the output and the input circuits becomes very high. For example, if the total grid-to-plate capacitance of a triode is 2 micromicrofarads, the reactance of this capacitance at 150 megacycles is approximately -530 ohms. Antiresonating this capacitance with an external coil of $+530$ ohms at 150 megacycles, and assuming a Q of 95, the grid-to-plate impedance would be raised to approximately 50,500 ohms. The result is a corresponding decoupling between the plate and grid circuits. This impedance is in series with the grid-to-cathode impedance and the plate-to-

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† Bell Telephone Laboratories, 463 West Street, New York, N. Y.

[1] United States No. 1,325,879, H. W. Nichols.

[2] United States No. 1,558,909, H. W. Nichols.

[3] A. L. Samuel and N. E. Sowers, "A power amplifier for ultra-high frequencies," PROC. I.R.E., vol. 24, pp. 1464–1483; November, 1936.

cathode impedance. From this relationship it follows that the magnitude of the voltage returned to the grid circuit from the plate circuit depends on the ratio of the grid-circuit impedance to the sum of the grid-to-plate impedance and the grid-circuit impedance. A similar relationship exists for the magnitude of the voltage fed through from grid to plate, in which, however, the plate-circuit impedance replaces the grid-circuit impedance. With zero plate voltage and a small radio-frequency voltage on the grid, the ratio of the grid voltage to the plate voltage is noted as the neutralizing circuit is brought into resonance and the grid and plate circuits are kept in tune. Following this procedure, a minimum value of the ratio of the plate voltage to the grid voltage may be obtained. This voltage ratio was observed by means of vacuum-tube voltmeters attached as closely as possible to the grid and plate elements of the coil-neutralized tube. Experimental observations verified the conditions expected, and there was realized in all of the models, for a good neutralizing condition, a feed-through plate voltage of approximately 4 or 5 per cent of the grid voltage.

When all three circuits had been brought simultaneously into tune, it was noted that almost complete isolation was effected between the grid and plate circuits so far as interaction on the grid circuit due to plate-circuit tuning was concerned. At this point, it is interesting to note that the equivalent capacitance between the grid and plate of a pentode or tetrode for a capacitive reactance of 50,000 ohms at 150 megacycles is 0.021 micromicrofarads, a value realized only in small commercial pentodes. The pentode or tetrode has the disadvantage of the screen-lead inductance, the reactance of which may seriously affect the satisfactory operation of the tube at very-high frequencies. Qualitatively, the decoupling achieved by coil neutralization is comparable with that obtainable with the best small pentodes, and, since it is applicable to tubes rated for transmitters, its advantages become more apparent.

Upon applying plate voltage to the neutralized triode, a change in the interelectrode impedances usually occurred, and vernier readjustments of all three circuits were required to obtain optimum neutralization as indicated by a minimum reaction on the grid voltage as the plate-circuit tuning capacitor was varied. Normal coupling to the plate-circuit load was maintained for both preliminary and final adjustments. The reaction noted above was but one third to one fourth the reaction usually observed with conventionally neutralized triodes, or with the available tetrode or pentode tubes when operated in this frequency range as single-side amplifiers.

With this qualitative discussion on the general properties of a coil-neutralized amplifier, our attention will now be turned to a description of the experimental apparatus used, and some phases of the experimental technique employed.

PART II. EXPERIMENTAL APPARATUS AND TECHNIQUE USED

Three push-pull coil-neutralized experimental amplifiers were built in succession using Western Electric 304A, 316A, and 226Y vacuum tubes. The performance of these amplifiers equaled, and, in some ways excelled, that of similar amplifiers using the bridge-capacitance type of neutralization. It was noted that the plate efficiency of the amplifiers was, to a large extent, a function of the frequency of operation, and the quality of the high-frequency design of the tube used. In this respect, efficiencies of 33 per cent were obtained with the 316A and the 226Y stages when operating as class B amplifiers at 150 megacycles. We will consider in detail the design of still another experimental model, as it is of greater technical interest.

With the development of the Western Electric 299Y triode,[4] which is capable of dissipating 100 watts under an air blast, a two-stage single-side coil-neutralized amplifier was constructed. Each tube was mounted in a removable frame on which the high-frequency circuits were built as compactly as possible. Simple shunt-resonant circuits were used in the grid, plate, and neutralizing circuits. Series-tuned circuits could be used, provided all three circuits were so designed and ganged together as to have the same type of reactance at any frequency, since this is the criterion for nonoscillation. In order more readily to effect neutralization, a vernier variable capacitor was placed between the grid and plate elements. This adjustment was made by means of a screw thread on a shaft which translated a metal plate relative to two fixed plates connected to the grid and plate elements respectively. Plate and grid diode voltmeters were integral parts of each unit. The power and measuring circuits were completed when the unit was seated in position, the two units being placed side by side in an aluminum box. Each tube was cooled by an air blast supplied by a blower through a duct in the rear of the assembly. Inductive coupling was used between the two stages, and between the second-stage plate circuit and the 320-ohm output load.

An exposed view of this two-stage model appears in Fig. 1. The bottom view of one unit appears in Fig. 2, in which the grid and plate circuits and diode voltmeters may be seen quite clearly on either side of a central longitudinal partition. The opposite side of this unit appears in Fig. 3, in which is shown the placement of the 299Y tube, and the coil and vernier tuning capacitor required in the neutralizing circuit. This tube, which is shown in Fig. 4, is provided with a pair of grid and plate terminals on diametrically opposite surfaces of the envelope, the advantages of which are apparent.

To set up two coil-neutralized amplifier stages in tandem, the first stage was neutralized as described above, using a high-frequency oscillator to supply the driving voltage. After neutralization, the plate power was applied and optimum neutralizing adjustments made. Using the excitation voltage provided from the first stage, the second stage was neutralized for minimum feed-through voltage when the load resistance was coupled to the plate circuit. Plate power was then applied and optimum output coupling effected. Readjustments were then made on the neutralizing control to obtain optimum neutralization. The direct-current bias voltage on each amplifier stage was adjusted to obtain a linear relationship between the driving voltage and the output voltage.

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[4] A. L. Samuel, "A negative grid triode oscillator and amplifier for ultra-high frequencies," PROC. I.R.E., vol. 25, pp. 1243–1252; October, 1937.

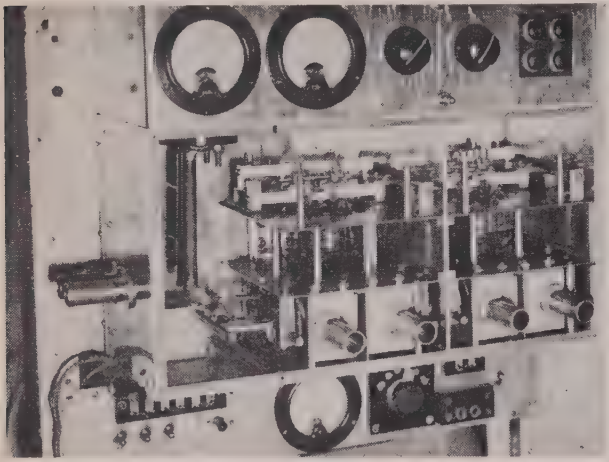


Fig. 1—Experimental two-stage coil-neutralized amplifier.

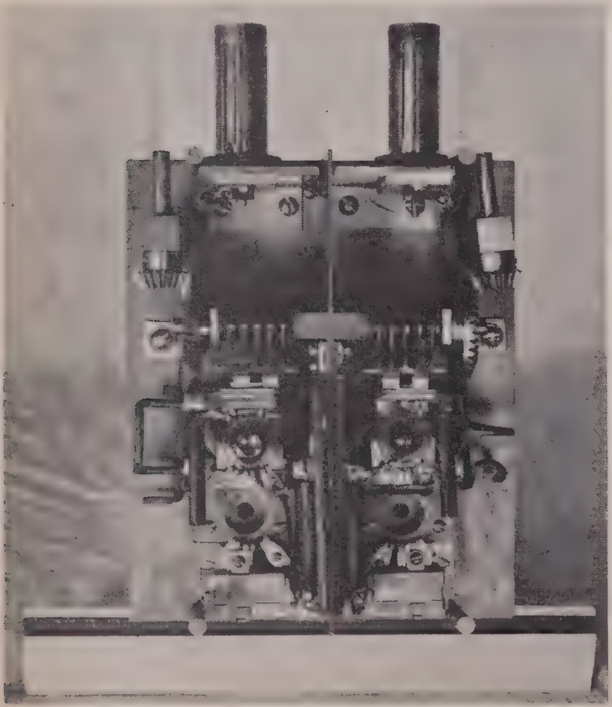


Fig. 2—Bottom view of a unit of the amplifier.

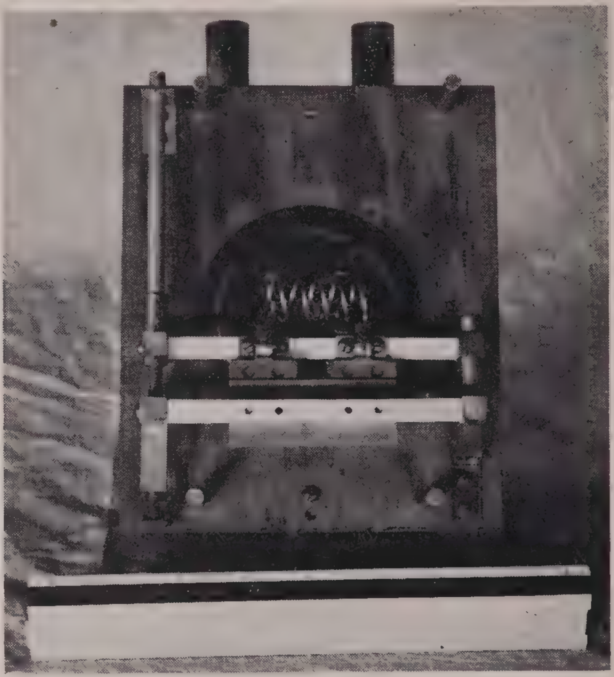


Fig. 3—Top view of a unit of the amplifier.

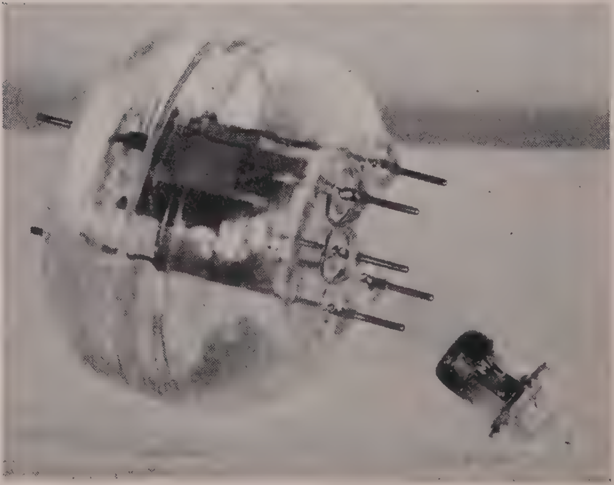
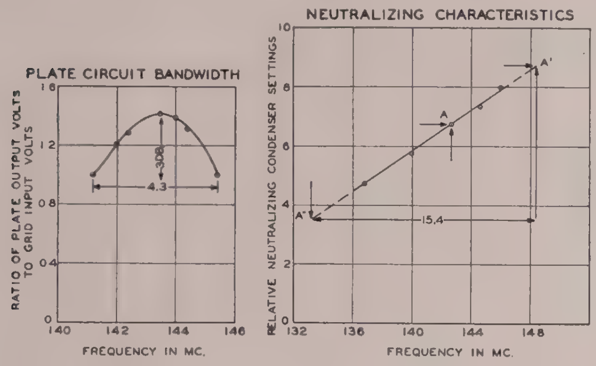


Fig. 4—The 299Y tube and the 955 monitor tube.

The second, or power-amplifier stage, is the more interesting of the two stages, since the first stage is merely a linear voltage amplifier which provides the necessary driving voltage. It was noted that in this stage the active grid shunting impedance[5] appeared to be many times greater than the impedance of the grid circuit, which was of the order of 10,000 ohms. This property is a measure of the effectiveness of the high-frequency design of the vacuum tube.

[5] M. J. O. Strutt and A. van der Ziel, "The causes for the increase of the admittances of modern high-frequency amplifier tubes on short waves," Proc. I.R.E., vol. 26, pp. 1011–1032; August, 1938.

The bandwidth of the output circuit when properly coupled to the load is of interest since this factor determines the extent of the frequency band which may be transmitted uniformly. This characteristic, given in Fig. 5(a), was obtained by applying a voltage of variable frequency but constant amplitude to the grid of the power tube and measuring the resulting voltage developed across the plate circuit. The grid and plate diode voltmeters were ideal for this purpose.



(a) (b)

Fig. 5—(a) Plate-circuit bandwidth characteristic of the power stage.
(b) Nonoscillating frequency interval of the power stage.

When testing an amplifier with a view to discovering its weaknesses or limitations as well as its strong points, it is good engineering practice to establish a criterion by which the stability of the amplifier may be described. In particular, the extent of maladjustment permitted in the neutralizing control before oscillations will occur is a useful guide to the stability of an amplifier. In bridge-capacitance neutralized circuits this latitude may be readily stated in terms of the micromicrofarads of capacitance variation permitted on either side of the optimum value of neutralizing capacitance. In the case of a coil-neutralized amplifier, the neutralizing is effected by the adjustment of a shunt-resonant circuit of high Q, and therefore a somewhat less direct method was used to obtain a criterion for the stability of the amplifier.

The technique of this method was as follows: The relationship between the vernier neutralizing-capacitor control and the frequency of operation of the amplifier was plotted as the operating frequency was varied from 135 to 150 megacycles. This relationship appears in Fig. 5(b). The amplifier was then set up to operate at 142.8 megacycles (point A on the curve of Fig. 5(b)). The excitation was removed and the neutralizing-capacitor control varied above and below the reference value until tuned-grid, tuned-plate oscillations occurred. These limits are indicated by the point A' and A'' on the curve. In this way, the relationship between the neutralizing-capacitor control range and the nonoscillating frequency interval was determined. Thus, a criterion for the stability of the amplifier may be established. We will now consider in detail typical operating characteristics of this coil-neutralized amplifier.

PART III. OPERATING CHARACTERISTICS OF THE COIL-NEUTRALIZED AMPLIFIER

Typical operating characteristics for the power stage operated as a class B amplifier at 144 megacycles are shown in Fig. 6. It is noted from curves D, B, and A that a useful output power of 65 watts was obtained at 31 per cent efficiency with a substantially linear characteristic between the input and the output voltage. The total power delivered by the tube is indicated by curve C. Some interest is attached to the measurement of power at this frequency and so the means employed will be briefly considered.

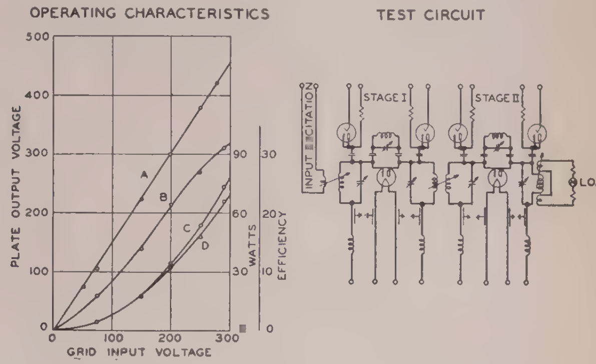


Fig. 6—Typical operating characteristics of the power stage. Curve A, input-voltage versus output-voltage linearity characteristic.
Curve B, useful-load power-efficiency characteristic.
Curve C, power delivered to the circuit and the load.
Curve D, useful power delivered to the 320-ohm load.

From the plate-circuit bandwidth determination (see Fig. 5(a)) the resonant step-up of the circuit Q is, to a good approximation, given by

$$Q = 144/4.3 = 33.5.$$

The plate-circuit inductance was determined to be

$$L = 0.043 \text{ microhenry.}$$

From these values, the antiresonant impedance of the loaded plate circuit was given by

$$Z = QwL = 1300 \text{ (ohms).}$$

The radio-frequency plate voltage was accurately given by the diode voltmeter. Therefore, the total power delivered by the tube as indicated by curve C is given by

$$W = (E^2/2)(1/Z)$$

where E is the diode peak voltage. The power delivered to the useful load was determined from the load current, and the value of the load resistance. The radio-frequency current-meter readings were corrected for the frequency of operation according to the manufacturer's calibration. In previous work it had been determined that the value of the load resistances was practically the same at this frequency as at power frequencies. These resistances were thin-walled isolantite tubes with a high temperature deposit of carbon on the external surface.

The power dissipated in the load was given by RI^2 and is indicated by curve D. The plate-circuit losses were given by the difference between curves C and D.

The curve of Fig. 5(b) indicates that a frequency interval of 15.4 megacycles was realized for the possible settings of the neutralizing control for which no oscillation would occur. About six tenths of this interval was below the operating frequency designated by point A on the curve. This appears to offer a reasonable margin of safety against maladjustments of the neutralizing-capacitor control. Independent variations of the plate and grid-circuit capacitors throughout their range indicated that no oscillatory condition could be produced from such adjustments.

No apparent increase in radio-frequency loading due to a shunting impedance resulting from electron transit-time effect in the tube was observed. There remained the possibility, however, that the not-readily observable variation of the transit angle,[6] which occurs during the radio-frequency cycle, might contribute to an increase in the normal nonlinear distortion generated in the amplifier. Should this effect be of significant proportions, it was expected that a systematic study of the distortion characteristics might reveal its contribution at least to a qualitative degree. In any event, the actual capabilities of the amplifier with regard to signal level, power output, and distortion, were of fundamental interest.

The amplifier brought to this stage of development appeared promising, and it became of interest to determine the distortion characteristics under conditions simulating those encountered in an actual communication system. The highlights of such a study and some typical results obtained will be presented in the following subdivision of this paper.

Part IV. Distortion Characteristics of the Coil-Neutralized Amplifier

In order to study as rigorously as possible the distortion characteristics of the coil-neutralized amplifier, it was decided to use a two-frequency generator and associated harmonic analyzer which had been developed with this purpose in view by associates working in the same frequency region. It will suffice to indicate the general principles used in this apparatus.

The twin-frequency generator consisted of two highly stabilized very-high-frequency oscillators of moderate power, differing in frequency by approximately 50 kilocycles, at a final frequency of approximately 144 megacycles. Great care was taken to prevent intermodulation of the two oscillators. This was realized to the extent that the dominant distortion products were 70 decibels below either fundamental frequency. Equal amplitudes were maintained for the two frequencies. The twin frequencies were supplied to the first circuit of the two stage coil-neutralized amplifier through a pair of shielded coaxial transmission lines.

The harmonic-distortion analyzer consisted of a high-frequency push-pull square-law detector with special input and intermediate-frequency circuits to provide the necessary selectivity to separate harmonic frequencies differing by but 50 kilocycles. An accurately calibrated, manually operated attenuator with a range of 100 decibels was used at the intermediate frequency in conjunction with a second-detector output meter also calibrated in decibels. Flexible shielded concentric-conductor transmission lines were used exclusively to connect the distortion analyzer to either amplifier stage through the short tubes evident in Fig. 1.

When two such pure frequencies are passed through a nonlinear device, such as a vacuum-tube amplifier, an intermodulation occurs which is evidenced by the appearance of a large number of new frequencies in the output which were not present in the input. This constitutes a distortion of the original input signal. We will be concerned only with the third- and fifth-order distortion products which arise in the following manner. Designate the two input frequencies by p and q. Then the dominant contributions to the third-order distortion in a reasonably linear amplifier are given by the terms $(2p-q)$ and $(2q-p)$. If, for example, p is 144.000 megacycles, and q is 144.050 megacycles, the third-order-distortion frequencies are 143.950 and 144.100 megacycles, respectively. These spurious frequencies appear but 50 kilocycles on each side of the pair of original frequencies.

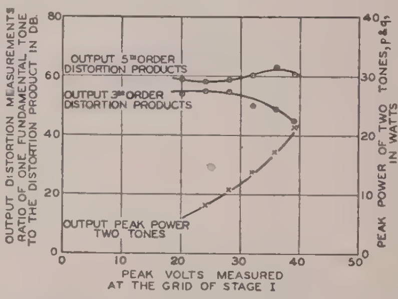


Fig. 7—Over-all distortion characteristics versus peak-power output. Two-tone distortion characteristics.
$p = 144.000$ megacycles
$q = 144.050$ megacycles
p and q are equal in amplitude.

In a similar manner, it can be shown that the dominant fifth-order products which result are given by the terms $(3p-2q)$ and $(3q-2p)$ which, for the given values of p and q, appear as frequencies of 143.900 and 144.150 megacycles, respectively. These are in turn removed but 50 kilocycles from the third-order terms. Still other orders of distortion products appear and add their terms at precise intervals of 50 kilocycles. The calibrated attenuator of the analyzer permitted the ratio of the distortion-product amplitude to fundamental amplitude to be obtained directly in decibels.

Using optimum distortion ratios as the criterion, the best value for the grid bias and for the inductive coupling between each tube and its load was determined, with the reservation that the minimum acceptable distortion ratio in the output of the second stage would be 40 decibels. This restriction limited the allowable power output of the second stage in a marked manner as would be expected. Typical results obtained are indicated in Fig. 7, and may be summarized as follows.

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[6] F. B. Llewellyn, "Phase angle of vacuum-tube transconductance at very high frequencies," Proc. I.R.E., vol. 22, pp. 947–956; August, 1934.

The peak-power output of the second stage occurred for the instantaneous coincidences of the maximum amplitude of the two input frequencies which were of equal amplitude, at which time, four times the power of a single frequency was developed. For a peak-power output of 23 watts, the third-order distortion-product amplitude was approximately 44 decibels below the amplitude of either fundamental. The fifth-order distortion-product amplitudes were better than 60 decibels below the amplitude of either fundamental. A voltage ratio of one hundred to one is the equivalent of a 40-decibel ratio in these measurements. In making distortion measurements, the monitoring diode filaments were turned off.

These observations on typical distortion characteristics, of which a number were made, seem to indicate that the dominant mechanism for the production of distortion may be attributed to the usual source; viz., the nonlinear properties of the tube, and that the effect of the electron transit time does not contribute appreciably to the distortion. With these results obtained, the objectives motivating the study of the coil-neutralized amplifier were, to a certain degree, attained, and while the data presented apply only to this model, there does not appear to be any reason to believe that the results given are not indicative of what may be realized with other tubes of similar type when operated in a similar manner.

Additional References

(1) United States No. 1,489,228, L. A. Hazeltine.
(2) United States No. 1,533,858, L. A. Hazeltine.
(3) United States No. 1,577,421, L. A. Hazeltine.
(4) United States No. 1,763,015, G. H. Stevenson.
(5) United States No. 1,832,261, G. H. Stevenson.
(6) W. H. Doherty, "Neutralization of radio frequency power amplifiers," *Pick-ups* (Western Electric Company), December, 1939.

Cathode-Follower Circuits[*]

KURT SCHLESINGER[†], ASSOCIATE, I.R.E.

***Summary*—The cathode follower is a circuit of increasing importance for television and ultra-high-frequency applications, and has therefore received much attention in technical literature. Most of the publications deal with the cathode-output amplifier, a three-terminal device. In this paper, such systems are treated briefly along with applications of the cathode follower for voltage amplification and signal mixing. This is preceded by a graphical diagram of the cathode-follower characteristic, which may be used for the determination of the direct-current components of such circuits.**

Cathode followers with a complex load are subsequently considered. These circuits show characteristic feedback phenomena from the cathode into the grid system in which the interelectrode capacitance acts as a coupling element. Most circuits of this nature are two-terminal devices, with action observed in the input circuit, and no external output derived from the cathode. The cathode-follower oscillator is perhaps the most important application of this group. Cathode-follower circuits with one or more resonant circuits in the cathode arm may be used as detectors or filters in connection with frequency modulation.

In presenting a variety of cathode-follower applications in the light of a general theory, this paper indicates the wide scope of possible applications with no intention of being exhaustive in every detail. In order to facilitate reading, most of the mathematical derivations have been put into the Appendix.

I. THE CATHODE FOLLOWER AS SIGNAL REPEATER

A. A Cathode-Follower Diagram

IN DESIGNING an amplifier circuit, the first step is to establish an appropriate working point along the direct-current characteristic of the tube. This demands the determination of adequate grid bias for given plate voltage and current. Fig. 1 shows a typical cathode-follower amplifier which derives grid bias from a tap along the cathode resistor. This technique has advantages as compared to bias from a bleeder across the B-supply, in that it tends to keep the plate current constant and relatively independent of plate-voltage variations.

Fig. 2 shows a graphical method of securing information about circuit design and signal transmission at low frequencies. This cathode-follower diagram is based upon a conventional triode characteristic, showing plate current versus grid voltage.

$$i_p = f(E_g). \tag{1}$$

This curve for a triode with grounded cathode and constant plate voltage E_b is plotted as c in Fig. 2.

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[*] Decimal classification: R139.21. Original manuscript received by the Institute, November 24, 1944; revised manuscript received, August 20, 1945.

[†] Columbia Broadcasting System, New York, N. Y. Work covered in this paper was done while the author was a member of RCA Laboratories, Purdue University, Lafayette, Indiana; material presented in a lecture, Purdue University, September 21, 1944.

The cathode-follower characteristic f for the circuit of Fig. 1 may be derived from this curve by reason of the fact that the plate current in a triode remains unchanged as long as the voltage sum $E_g+(1/\mu)E_p$ is kept constant. In this expression, E_g and E_p are voltages measured between the tube electrodes. This expression may be rewritten in terms of input and output voltages, e_g and e_k, respectively:

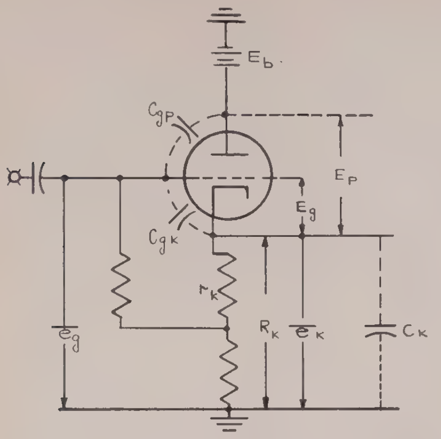

Fig. 1—Typical cathode-follower amplifier.

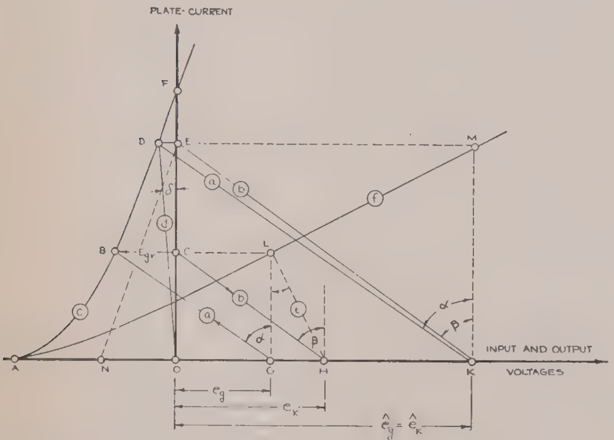

Fig. 2—Cathode-follower characteristic. Angle GLH$=\gamma$.

$$E_g = e_g - e_k \tag{2}$$

$$E_p = E_b - e_k \tag{3}$$

$$E_g + \frac{1}{\mu}E_p = e_g - e_k\left(1+\frac{1}{\mu}\right) + E_b/\mu. \tag{4}$$

The same plate current,

$$i_p = e_k/R_k \tag{5}$$

is then found to flow under the influence of a plate voltage E_b and a grid bias of

$$E_g = e_g - i_pR_k\left(1+\frac{1}{\mu}\right). \tag{6}$$

In Fig. 2, the plate current may then be read from the point of intersection **B**, where the characteristic c, given by equation (1), meets the load line a, given by equation (6).

In this way, the output voltage $e_k=$**OH** may be obtained for each input voltage $e_g=$**OG** by constructing the quadrilateral $GBCH$ with **BC** parallel to the abscissa and with the load lines a and b having slopes, respectively, of

$$\tan\alpha = R_k\left(1+\frac{1}{\mu}\right) \tag{7}$$

$$\tan\beta = R_k. \tag{8}$$

The actual cathode-follower characteristic f showing the plate current **OC**$=$**GL** versus input voltage **OG**$=$**CL** may then be constructed point by point. If the working point **L** is to be established by a cathode-resistor tap, r_k, Fig. 1, its value may be deduced from the angle $\gamma=GLH$

$$\tan\gamma = r_k \tag{9}$$

according to equation (2).

From Fig. 2 it is seen that the cathode-to-ground voltage **OH** is more positive than the grid-to-ground voltage for all working points below **M**. At this point, grid current begins to flow.

To find the peak plate current, **OE**$=i_{p(\max)}$, a load line d may be drawn through the origin O at the angle δ

$$\tan\delta = \tan\alpha - \tan\beta = R_k/\mu. \tag{10}$$

The peak output voltage **OK**$=i_{p(\max)}\cdot R_k$ is then found to be μ times **DE**, which indicates the desirability of using a large cathode resistor R_k.

As a by-product of these considerations, a convenient method of finding amplification factors is available. From inspection of the triangles ODE and KDM, it is apparent that the same plate current will flow in a triode which has: (a) the grid at ground and r_1 ohms between cathode and ground; or (b) grid and cathode connected and r_2 ohms between cathode and ground. The amplification factor for that plate-current value is then given by the equation

$$\mu = r_2/r_1 - 1. \tag{11}$$

B. A Multistage System for Signal Synthesis

The preceding section has shown that the signal output from a cathode follower differs from the input in two ways: (1) by a slight reduction in amplitude, and (2) by the addition of a positive-voltage bias. These properties may be used in signal transmission systems where it is desired to combine a plurality of signals at definite amplitude levels.

A typical problem of that kind occurs in television transmission where synchronizing pulses are to be added to the picture signal. Shortly before these pulses are transmitted, the signal is supposed to reach the black level and to hold it for some time, regardless of the instantaneous level of the video modulation. This problem may be solved by the use of several cathode-follower tetrodes in cascade, as shown schematically in Fig. 3. This circuit functions as a mixer for video, direct-current, blanking, and synchronizing signals.[1] Each stage has its inner grid directly connected to the cathode of the preceding stage, while negative pulses are applied to the screen grids from the blanking and synchronizing-pulse generators 7 and 8, respectively.

The graphs (*a*) through (*f*) in Fig. 3 explain the synthesis of the final television signal, step by step. Graph (*a*) shows the alternating-current component of the video signal *V*, obtained from the pickup tube 1, and the video amplifier 2. This signal reaches a direct-current restorer 3, through a coupling capacitor, and leaves it with the negative peaks raised to zero level, as shown in graph (*b*) of Fig. 3. Graph (*c*) illustrates the function of the first keying tetrode 5. The output from this stage stays above zero level even at those instants when the signal input is zero. The output from stage 5 reaches zero level, however, if negative pulses arrive at the screen grid and key the plate current. This brings the signal to a definite level or blanking pedestal, as shown in graph (*d*) of Fig. 3, regardless of the video voltage at the time. This action is repeated in the second cathode-follower tetrode 6, and is explained by another application of the cathode-follower diagram shown is graph (*e*), with the resulting signal output of graph (*f*). In this final form, the television signal exhibits synchronizing pulses of the amplitude *S* superimposed on a constant black level *B* which is maintained throughout the transmission. In spite of the loss in signal amplitude at each stage, the signal output is of the same order as the input because each stage adds a pulse component to the original signal. There is no particular difficulty in transmitting a wide band of frequencies through the cathode-follower cascade since the output impedance of each stage is sufficiently low.

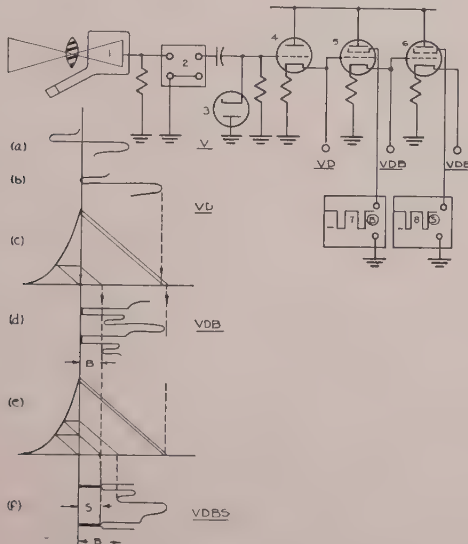

Fig. 3—Cathode-follower cascade for television transmission.

II. CATHODE-OUTPUT AMPLIFIERS

As long as there is no appreciable phase lag between cathode and grid, the cathode follower exhibits low input admittance and low output impedance. These qualities are utilized in well-known amplifier applications, particularly in connection with video amplifiers. In such circuits a cathode-follower stage is commonly used either to match the high output impedance of a television pickup tube,[2] or to match the low input impedance of a cable.[3–5] Cathode-follower action is frequently employed to correct falling frequency response of video amplifiers and to expand their bandwidth (Section II, C). The cathode follower may also be used as an interstage coupling element in video amplifiers.[6,7] Although the gain per stage may be more than doubled by this method, such a circuit is not widely used, because it is less economical than modern methods of network coupling. Another cathode-output amplifier which yields a gain greater than unity consists of a step-up transformer at the output of a cathode follower (Section II, D). This combination may be used in audio amplifiers.

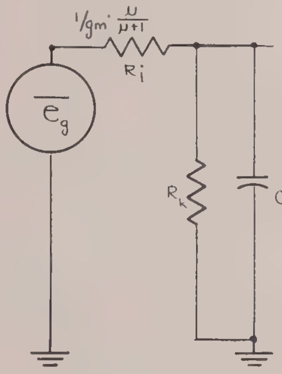

Fig. 4—Equivalent circuit of cathode-output amplifier of Fig. 1. Generator electromotive force, shown as $\bar{e}_g$, is equal to $e_g \cdot \mu/\mu+1$.

A. Gain and Bandwidth

The behavior of a cathode-output amplifier, such as Fig. 1, may be readily deduced from the equivalent circuit of Fig. 4. This shows the triode as a generator with the electromotive force $e_g\mu/(\mu+1)$ and with the source impedance[8]

$$R_i = 1/g_m \frac{\mu}{\mu + 1}. \tag{12}$$

From Thevenin's theorem, it follows that the gain and

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[1] The circuit description and diagram, Fig. 3, are simplified to explain the general idea of operation. Further details regarding a practical form of the circuit may be found in U. S. Patent No. 2,366, 358, January 2, 1945, issued to Kurt Schlesinger, assignor to RCA.

[2] A. A. Barco, "Iconoscope preamplifier," *RCA Rev.*, vol. 4, pp. 89–107; July, 1939.

[3] Albert Preisman, "Some notes on video amplifier design," *RCA Rev.*, vol. 2, pp. 421–432; April, 1938.

[4] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Company, New York, N. Y., 1943, p. 431.

[5] V. K. Zworykin and G. A. Morton, "Television," John Wiley and Sons, New York, N. Y., third edition, pp. 438–439.

[6] J. G. Brainerd, G. Koehler, Herbert J. Reich, and L. F. Woodruff, "Ultra-High-Frequency Techniques;" D. van Nostrand Company, New York, N. Y., 1942, pp. 221–223.

[7] H. J. Reich, "Features of cathode-follower amplifiers," *Elec. Ind.*, vol. 4, pp. 74–77; July, 1945.

[8] See Appendix I.

TABLE I

Tube Type	Amplification Factor	Transconductance g_m (micromhos)	Load		Cutoff Frequencies		Total Bandwidth b eq. (18) megacycles	Gain a eq. (17)	Maximum Signal Output for 10-Megacycle Bandwidth and 10-Milliampere Plate Current e_k eq. (21) Volts
			Capacitance C_k (microfarads)	Resistance R_k eq. (19) (ohms)	Internal f_i eq. (15) (megacycles)	External f_e eq. (16) (megacycles)			
6J5	20	3000	16	10,000	30	1.0	31.0	0.92	9.95
			80	497	6	4.0	10.0	0.57	9.85
			160	143	3	7.0	10.0	0.28	0.82
6AC7	50	8000	16	10,000	80	1.0	81.0	0.97	9.95
			80	10,000	16	0.2	16.2	0.96	2.0
			160	500	8	2.0	10.0	0.78	0.98
6J4	55	12,000	16	10,000	120	1.0	121.0	0.97	9.95
			80	10,000	24	0.2	24.2	0.91	9.95
			160	10,000	12	0.1	13.0	0.91	9.95

frequency response of the cathode follower are the same as that of an equivalent pentode with grounded cathode which has the same load capacitance C_k and a plate resistor of

$$r_p = \frac{1}{g_m(1 + 1/\mu) + 1/R_k} \text{ ohms.} \tag{13}$$

This is the value of the parallel connection of source resistance R_i (equation (12)) and load resistance R_k. The angular frequency at which the gain of the equivalent amplifier falls off 3 decibels from its value for low frequencies is the reciprocal of the time constant r_pC_p of the equivalent plate circuit.

The bandwidth of a cathode-output amplifier may thus be found from equation (13).

$$b = \frac{g_m(1 + 1/\mu)}{2\pi C_k} + \frac{1}{2\pi C_k R_k}. \tag{14}$$

In this form, the cathode-follower bandwidth is seen to be the sum of two separate cutoff frequencies

$$\frac{g_m(1 + 1/\mu)}{2\pi C_k} = f_i \tag{15}$$

$$1/2\pi R_k C_k = f_e. \tag{16}$$

These may be termed "internal" and "external" cutoff frequencies because, with a given load capacitance, the first depends on the parameter g_m of the tube, and the second depends on the resistor R_k in the external circuit. In these terms the amplification factor becomes

$$a = \frac{f_i}{f_i + f_e} \cdot \frac{\mu}{\mu + 1} \tag{17}$$

and this gain will be 3 decibels down at the bandwidth

$$b = f_i + f_e. \tag{18}$$

In many cases this bandwidth is more than adequate to cover the spectrum f_s of signal frequencies. (See Table I.) If this is not true in a given case, however, the cathode-follower bandwidth may be expanded by increasing the external cutoff frequency. This is done by decreasing the value of the cathode resistor according to the equation

$$R_k \leqq 1/g_m \cdot \frac{f_i}{f_s - f_i} \quad \text{for} \quad f_s > f_i. \tag{19}$$

This extension of the bandwidth is achieved, however, at the expense of amplification.

Multiplying (17) and (18) there results

$$a \cdot b = f_i \cdot \frac{\mu}{\mu + 1} = \text{constant} \tag{20}$$

which indicates that the product of gain and bandwidth remains constant. A similar rule is known to hold for plate-loaded amplifiers, with the difference, however, that in the case of the cathode follower, a reduction of the signal band below the internal cutoff frequency (equation (15)) cannot be exchanged for amplification factors greater than one.

In Table I, the internal cutoff frequencies are listed for tubes with medium and high transconductance and for large and small load capacitances. The cathode-resistor values are computed for a signal bandwidth of $f_s = 10$ megacycles.

It is possible to compensate cathode-output amplifiers at high frequencies by methods similar to those used for plate-loaded amplifiers. At the cathode-follower output, series peaking is quite successful. Fig. 5 shows such a peaking coil L_s, with shunt resistor which is frequently needed to avoid overshoot. No such damping resistor is required if the peaking inductance is kept small enough so that the circuit composed[9] of L_s, C_g, and $1/g_m$, has a Q factor of less than 0.6. This condition is satisfied if series resonance between L_s and C_g occurs at or above a frequency of 1.5 times the cutoff frequency. If this

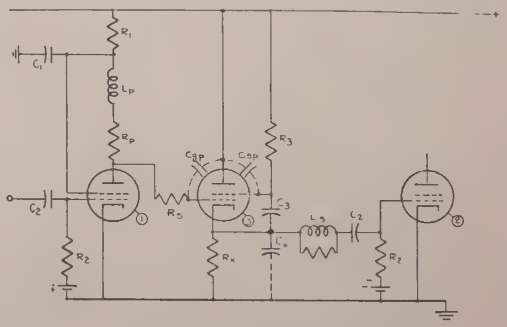

Fig. 5—Cathode-follower tetrode as an interstage coupling tube.

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[9] Heinz E. Kallman, Rolf E. Spencer, and C. P. Singer, "Transient response," PROC. I.R.E., vol. 33, pp. 169–195; March, 1945.

is done, the cathode follower is compensated over the entire bandwidth and shows good transient response.

In the cathode-follower amplifiers, only a portion of the cathode bias voltage is available as signal output over a given bandwidth. Unlike the plate-loaded amplifier with low plate impedance, a cathode follower may be designed with a high cathode impedance R_k, and still be able to cover a wide band of frequencies, including the internal cutoff frequency. In this design, however, larger amplitudes may be handled at low frequencies than at high frequencies, with a given plate current and without distortion. With a system as shown in Fig. 1, the peak-to-peak output voltage $|2e_k|$, available up to a frequency f_s, may be written in terms of the crest plate current, I_o, which is represented as ***OE*** in Fig. 2.

$$|2e_k| = \frac{I_oR_k}{\sqrt{1+(f_s/f_e)^2}} < I_oR_k\cdot\frac{f_e}{f_s}\cdot \qquad (21)$$

This output swing obviously decreases with increasing frequency. If the cathode follower is to be used over its entire bandwidth (equations (14) and (18)), the signal swing reduces to

$$|2e_k| = \frac{I_o}{g_m + 1/R_k} < \frac{I_o}{g_m}\cdot \qquad (22)$$

This is represented in Fig. 2 as the distance ***ON***. Measurement of frequency response in cathode-follower circuits may be in error if made with amplitudes in excess of those indicated. Moreover, the low source impedance of the cathode-follower cannot be guaranteed for amplitudes beyond those limits. If this output is too small, several tubes have to be connected in parallel, in order to increase the crest current I_o.

B. Interstage Coupling by Cathode Followers

In Fig. 5, a tetrode is used as a coupling element in cathode-follower connection. The screen grid of this stage is connected to the cathode through by-pass capacitor C_3, rather than to ground. This capacitor is so designed that its susceptance matches the tube transconductance in midband. It then prevents low frequencies from the plate supply from reaching the output, but it connects screen and cathode in parallel for the higher signal frequencies. As a result, the grid-to-plate capacitance C_{gp} of the coupling tube is effectively removed from the plate output of the driver stage *1*. In return, an additional load capacitance C_{sp} appears across the cathode output, but this is readily disposed of by the low output impedance of the cathode follower.[10] A tetrode as coupling tube yields about 24 per cent more gain at the higher frequencies than a triode connection of the same tube. This gain is obtained, however, at the expense of a somewhat higher noise level, due to random current distribution between screen grid and plate.[11,12]

The plate of the driver stage 1 may be directly connected to the grid of the coupling tube 3, provided that sufficient voltage drop is established across the filter resistor R_1. This leaves the circuit with only two resistance-capacitance systems of opposite frequency response between which low-frequency correction may be obtained ($R_pC_1=R_2C_2$). High-frequency compensation may be accomplished by a series inductance L_s at the output of the coupling stage. In addition to this, the input signal may be compensated by a shunt peaking coil L_p in the plate circuit of the driver stage. The grid resistor r_s in Fig. 5 acts as a feedback-suppressor without which regeneration is likely to occur, due to the negative input resistance of the cathode-follower coupling stage with capacitive load. (See Section III.)

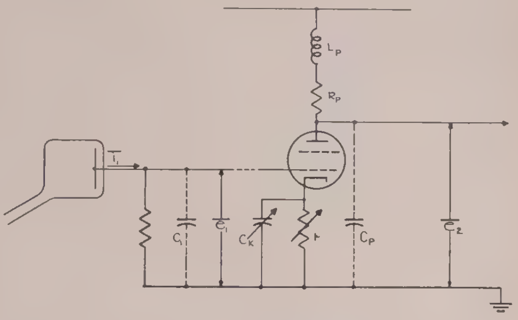

Fig. 6—Frequency-band expansion by cathode-follower amplifier.

Voltage-amplification factors of 30:1 may be obtained over a 6-megacycle band by using this circuit arrangement with 6AC7 tubes as amplifier and coupling tetrode.

C. Frequency-Band Expansion by Cathode-Follower Action

Fig. 6 shows another application of the cathode follower for video amplification. In this case, the cathode-to-ground voltage is not connected to any external output; instead, it is used to modify, in a predetermined way, the grid control in an equalizer stage. The practical problem consists in equalizing over a wide band f_2 a signal input $\bar{e}_1$ with insufficient amplitude at high frequencies. Suppose the signal is obtained across an input circuit r_1C_1 from a constant-current generator. The input voltage,

$$\bar{e}_1 = \frac{\bar{i}_1r_1}{1+pr_1c_1}; \qquad p = j\omega, \qquad (23)$$

exhibits a drop of 3 decibels at the bandwidth f_1

$$f_1 = 1/2\pi r_1C_1. \qquad (24)$$

It is desired to expand this band s times

$$s = f_2/f_1. \qquad (25)$$

In the equalizer stage with cathode degeneration, the plate current becomes

$$i_p=g_m(\bar{e}_1-\bar{e}_k)=\frac{\bar{e}_1g_m}{1+\dfrac{g_mr_k}{1+pr_kC_k}}=\bar{e}_1g_m\frac{1+pr_kC_k}{1+p\dfrac{r_kC_k}{1+g_mr_k}}\cdot \qquad (26)$$

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[10] See Appendix I, equation (12).

[11] D. O. North, "Fluctuations in space-charge currents," *RCA Rev.*, vol. 5, pp. 244–260; October, 1940.

[12] W. A. Harris, "Fluctuations in space-charge currents," *RCA Rev.*, vol. 5, pp. 505–524; April, 1941.

By combination of (26) and (23) and multiplication by the plate impedance, the over-all amplification of the system is obtained.

$$\frac{e_2}{i_1 r_1} = \frac{g_m r_p}{1+g_m r_k} \cdot \frac{1+pr_k C_k}{1+pr_1 C_1} \cdot \frac{1+pL_p/r_p}{1+pr_k C_k/(1+g_m r_k)}. \tag{27}$$

From this expression it may be seen that the bandwidth over which constant gain is available may be expanded *s* times (equation (25)) if the following three conditions are fulfilled:

$$r_k C_k = r_1 C_1 \tag{28a}$$

$$1 + g_m r_k = s \tag{28b}$$

$$L_p/r_p = 1/2\pi f_2. \tag{28c}$$

The first two equations define a cathode circuit in which both the resistance r_k and the by-pass capacitance C_k are unique[13]

$$\begin{aligned} r_k &= 1/g_m \cdot (s-1) \\ C_k &= g_m/2\pi f_1(s-1). \end{aligned} \tag{29}$$

With no plate inductance L_p, the output will then be 3 decibels down at the frequency f_2. Perfect compensation requires a plate circuit with "peaking coil" and with a time constant as specified by equation (28c). It may be shown that, in this design, the plate resistor must be limited to

$$r_p = 1/4\pi f_2 C_p \tag{30}$$

in order to avoid overshoot. This is one half of the resistance value for conventional amplifiers with shunt peaking. The over-all gain is then

$$a = \frac{e_2}{i_1 r_1} = \frac{g_m r_p}{s} \tag{31}$$

so that the product of gain by bandwidth is once more found to be constant. This method of bandwidth expansion by cathode-follower action has been used with success in amplifiers for television pickup tubes.[14]

D. Cathode-Follower Amplifier with Output Transformer

If the load resistance is considerably higher than the output impedance of the cathode follower, voltage amplification may be obtained by the use of a step-up transformer between the cathode output and the load. Such a circuit, as shown in Fig. 7, may be used as a driver for audio amplifiers. Throughout the audio bandwidth of 10,000 cycles, a gain of 7:1 may be obtained with tubes of high transconductance, and a gain of 4:1 with ordinary triodes of the type[15] 6J5.

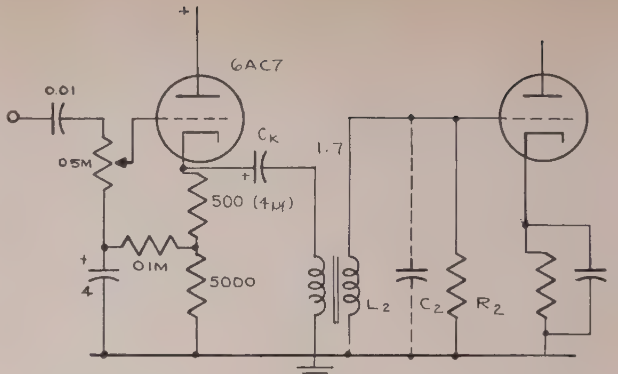

Fig. 7—Cathode-follower audio amplifier with a step-up transformer.

The over-all fidelity of this circuit may be much improved by tuning the primary of the transformer to a low frequency, using an appropriate coupling capacitor C_k. The low output impedance of the cathode follower facilitates series resonance between the coupling capacitor and the transformer primary. This effect may be used to emphasize the low audio frequencies, or to extend the transmission band toward very low frequencies. Fig. 8 shows the influence of various values of the coupling capacitance on low-frequency response. The high-frequency peak is due to resonance in the transformer secondary.

The circuit of Fig. 7 has less gain than plate-loaded amplifiers,[16] but it offers a number of advantages. Since it is a negative feedback circuit, it shows little harmonic distortion even at large amplitudes. Transformer distortion is also reduced because direct current is blocked from the primary. Furthermore, the cathode follower acts as an efficient filter[17] to remove the hum and low-frequency fluctuations from the power supply.

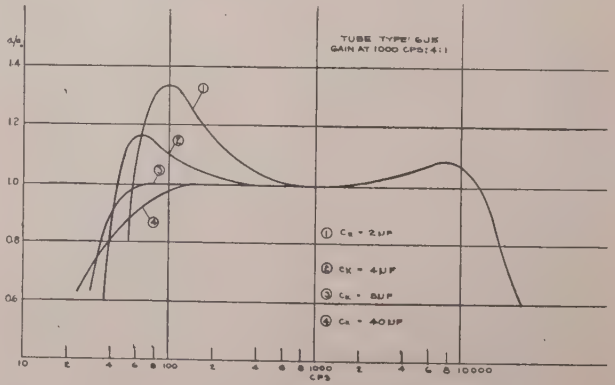

Fig. 8—Low-frequency response of the circuit of Fig. 7.

III. Two-Terminal Cathode-Follower Devices

A. Complex Cathode Impedance

Cathode-follower systems with a complex cathode impedance may be called complex cathode followers. In such systems, cathode and grid voltages may be considerably out of phase, and characteristic feedback phenomena occur through the grid-cathode capacitance as a coupling element. Fig. 9 shows three prototypes of such circuits. Circuits *a* and *b* each have only one reactance element in the cathode arm; it is negative in circuit *a* and positive in circuit *b*. Circuit *c* has two cathode reactances of opposite sign, forming a series-resonant circuit. The total cathode reactance then becomes a function of frequency and may change from negative values below resonance to positive values above resonance. More complex cathode networks with three or more reactances may be reduced to one of these prototypes over limited frequency ranges.[18] A study of these prototypes therefore gives a fairly general picture of what may be obtained from such circuits in actual practice.

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[13] Due to the current distribution between plate and screen grid, the transconductance of a pentode is 10 to 20 per cent higher in the cathode circuit than it is in the plate circuit.

[14] U. S. Patent No. 2-378-797, issued to Otto H. Schade, assignor to RCA, June 19, 1945. See also U. S. Patent No. 2-384-263.

[15] Derivation, see Appendix.

[16] The gain inferiority is of the order of $1/\sqrt{\mu}$.

[17] Filtering factor $=1/\mu+1$.

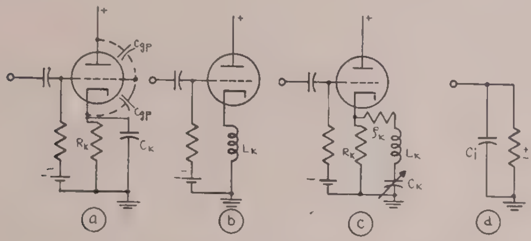

Fig. 9—Cathode-follower with complex load.

Analysis[19] shows that in all three cases the input impedance of the complex cathode follower appears to be a capacitance C_i shunted by a resistance R_i (circuit *d*, Fig. 9). This general behavior of the complex cathode follower is illustrated by the vector diagram of Fig. 10.

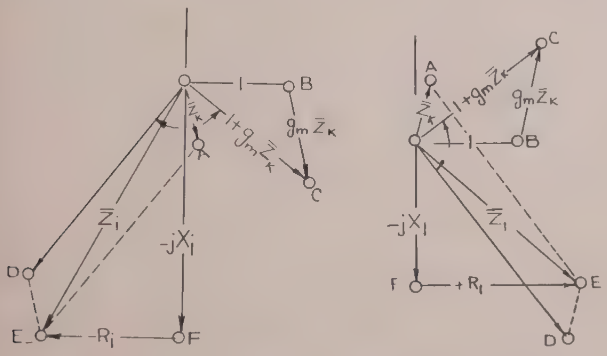

Fig. 10—Vector diagram of complex cathode follower.

It shows how the input impedance **OE** may be derived from a given cathode load vector **OA**. **OB** is the real unit, and **OD** is obtained by a clockwise rotation of 90 degrees from **OC** as a base. To obtain the input impedance vector **OE**, the cathode load **DE** is added to **OD**. In Fig. 10 is shown a capacitive-resistive cathode load, and an inductive-resistive cathode load with equal impedance and opposite phase angle. The resulting input impedances both have a capacitive-reactance **OF** and almost equal resistive components **EF**. However, the polarity of the input resistance is negative for capacitive load and positive for inductive load. Accordingly, degeneration is caused by inductance and regeneration by capacitance in the cathode branch.

More information is obtained from Fig. 11, which shows both input parameters C_i and R_i as functions of the cathode reactance for a complex cathode follower with tuned-cathode circuit, such as circuit *c* of Fig. 9. For such systems, the cathode reactance may be written

$$X_k = \sqrt{\frac{L_k}{C_k}}\,(x - 1/x); \quad x = f/f_{\text{res}} \tag{32}$$

where x is the detuning factor. It is seen that the cathode reactance near resonance ($x=1$) is proportional to the frequency deviation from resonance, so that the abscissa g_mX_k may be interpreted as a frequency scale.

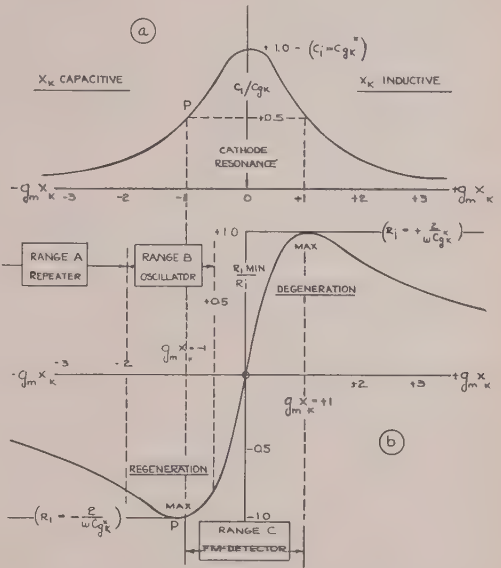

Fig. 11—Input parameters of the complex cathode follower as functions of frequency.

At resonance, the input capacitance reaches a maximum equal to the grid-cathode capacitance. The frequency deviations on either side of resonance have relatively little influence, and the capacitance variations are almost symmetrical. On the other hand, the resistive component of input admittance changes sign as the frequency passes through cathode resonance. To the left of $X_k=0$; (i.e., frequencies below resonance), the input conductance becomes negative, and regeneration results. To the right of $X_k=0$; (i.e., for frequencies above cathode resonance) there is degeneration in the input circuit.

The feedback actions increase on either side of cathode resonance, and reach their maxima for two frequencies, ω_1 and ω_2, where[20]

$$g_mX_k = \pm 1. \tag{33}$$

At these points the input conductance assumes a positive or negative maximum. Analysis shows that the minimum value of the input resistance amounts to twice the reactance of the grid-cathode capacitance at those frequencies.

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[18] T. E. Shea, "Transmission Networks and wave filters," D. van Nostrand Company, New York, N. Y., 1929; part 1, chapter 5, pp. 124–167.

[19] See Appendix III, equation (53).

[20] In (33) and (34) it is assumed that $\omega Cg < g_m$, which excludes very high frequencies. For more complete expressions see Appendix III.

$$|R_i|_{\min} = \frac{2}{\omega_{\frac{1}{2}} C_{gk}}. \tag{34}$$

For greater frequency deviations from cathode resonance, the feedback action decreases again. Equation (34) is important for the design of cathode-follower oscillators.

The technical applications of the complex cathode follower may be broadly classified into three groups: signal repeater, oscillator, and detector for frequency modulation. These are shown in Fig. 10, as ranges *A*, *B*, and *C*, respectively, along the frequency scale.

It was shown in Section II, *A*, that repeater action occupies a bandwidth from zero frequency up to the internal cutoff frequency. Low input capacitance and negligible feedback into the grid circuit are characteristic for this mode of operation. Range *A*, or the repeater range, is indicated as including all negative values of cathode reactance up to approximately $g_m X_k = -2$, where repeater action gives way to oscillator action. Range *B* is represented for oscillator applications. It lies on either side of the point *P*, where the input conductance reaches its negative peak. The third range, *C*, is centered around cathode resonance. In this range, the feedback into the grid circuit alternates from regeneration for positive frequency excursions to degeneration for negative frequency excursions. By reason of this fact, the complex cathode follower may be used as a frequency discriminator; some such applications will be described in Sections III, *C*, and IV, of this paper.

B. The Oscillating Cathode Follower

It was shown in the preceding section that the complex cathode follower with a capacitive load has negative input resistance. Analysis and experience lead therefore, to the expectation that oscillations may be generated in a tuned circuit connected between grid and ground, such as that of Fig. 12. By diagramming the electrode capacitance C_{gk}, the circuit becomes similar to a Colpitts-type oscillator. Analysis of this oscillator by methods developed in the theory of the complex cathode follower yields valuable data regarding design factors and performance.

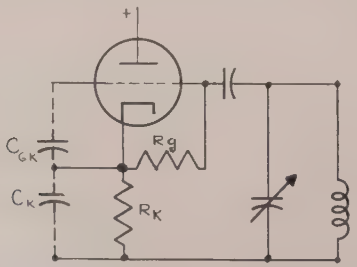

Fig. 12—Cathode-follower oscillator system. Note that tuned circuit is connected between grid and ground.

The system of Fig. 12 has a negative cathode reactance

$$X_k = -1/\omega C_k. \tag{35}$$

There is practically no equivalent resistance in series with X_k, especially if chokes are employed in the cathode supply. Under these conditions, the system exhibits a negative input resistance[21] of

$$R_i = -\frac{1}{2\pi f}\left[\frac{1}{C_g} + \frac{1}{C_k}\right] \cdot \left[\frac{f}{f_0} + \frac{f_0}{f}\right]. \tag{36}$$

It is evident from this equation that the input conductance reaches a negative peak value at a frequency f_0 which is found as

$$f_0 = \frac{g_m}{2\pi[C_g + C_k]}. \tag{37}$$

At this preferred frequency, the input resistance becomes

$$R_{i0} = \frac{-1}{\pi f_0}\left[\frac{1}{C_g} + \frac{1}{C_k}\right]. \tag{36a}$$

This is just twice the value of the reactance offered by the capacitive voltage divider between grid and ground.

At any frequency f other than the preferred one f_0, the input resistance is obtained by multiplying R_{i0} by the factor

$$R_i/R_{i0} = \tfrac{1}{2}[f/f_0 + f_0/f].$$

By combining equations (36) and (37), the minimum input resistance may also be written in the form

$$R_{i0} = -\frac{2}{g_m}\,\frac{(C_g + C_k)^2}{C_g C_k}. \tag{38}$$

This yields the important result that an optimum of design is obtained if the capacitances on either side of the cathode are made equal, $C_g = C_k$. If this is done, the negative input resistance is reduced to its lowest possible value

$$Ri_{\min} = -\frac{8}{g_m}. \tag{38a}$$

This may be realized at any frequency f, provided that the coupling capacitances are chosen as follows:

$$C_g = C_k = \frac{g_m}{4\pi f}. \tag{39}$$

The result of this theory may be summarized as follows: *The cathode-follower oscillator has a preferred range of operation at or around a frequency where the susceptances of the cathode-to-ground and cathode-to-grid capacitances combined are equal to the transconductance of the tube. At this frequency, the negative input resistance of the oscillator assumes a minimum value equal to twice the capacitive reactance between grid and ground. The maximum of the negative input conductance is 1/8 of the transconductance of the tube. This is obtained at the preferred frequency and with equal capacitances on either side of the cathode.*

This theory answers one or both of the following questions: (1) What is the best load capacitance for oscillator operation at some predetermined frequency? (2) Which

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[21] Derivation given in Appendix III, equation (54).

TABLE II

Type	Transconductance g_m (mhos)	Grid-Cathode C_g (micromicro-farads)	Cathode-Ground C_k (micromicro-farads)	Preferred Frequency f_0 (megacycles) eq. (37)	Input Resistance at Preferred Frequency R_{i0} (ohms) eq. (36a)	Smallest Possible Input Resistance $R_{i\,opt}$ eq. (38a)	Input Resistance at Frequencies of 5 mega-cycles	50 mega-cycles	500 mega-cycles
6J5	3000	3.0	8.0	43.5	−3070	−2670	−13,500	−3370	−17,700
6AC7	9000	11.0	9.0	71.5	− 900	− 890	−6460	−960	−3230
6AG7	11,000	5.2	10.7	110.0	− 830	− 726	−9000	−1100	−1980
6F4	5800	2.0	3.0	185.0	−1440	−1380	−26,600	−2850	−2200
6J4	12,000	5.5	2.8	230.0	− 750	− 667	−17,300	−1810	− 990

is the preferred frequency range for an oscillating cathode follower with a given load? This latter question is of particular interest if the tube is to oscillate with only its natural cathode-to-ground capacitance as load.

In Table II, numerical data are listed for cathode-follower oscillators at various frequencies and for tubes of medium and high transconductance. It is assumed that there is no feedback path other than the interelectrode capacitances. Both the negative input conductance and the tendency to oscillate increase with increasing frequency until the preferred range is reached. At broadcast frequencies the input resistance keeps well above −100,000 ohms. This is hardly sufficient to start oscillations in practical circuits with losses. It becomes necessary, therefore, to connect additional external capacitance between grid and cathode. As the frequency increases, the input resistance drops until values as low as −10,000 ohms are available at 10 megacycles, and less than −1000 ohms at 100 megacycles and above. At the same time, however, the optimum cathode-to-ground capacitance decreases to a value which eventually become comparable to, or even smaller than, the cathode-electrode capacitance itself. (See equation (39).) In designing cathode-follower oscillators for still higher frequencies, the cathode load capacitances for optimum design become so small that a serious mismatch cannot be avoided. Fig. 13 presents a balanced cathode follower oscillator for ultra-high frequencies which is particularly suited for demonstration purposes. It has two 6AG7's in triode connection, and a total plate input of about 10 watts. The Lecher-wire system, (parts 1 and 2), is connected to a coupling rod 3, about 1 inch long, which is grounded at the center tap 4. The plates of the tubes are grounded through the plate battery 8 and by-pass capacitors, while the cathodes follow the grid oscillations with lagging phase and are loaded by their natural electrode capacitance only.[22] Feedback occurs solely through the grid-cathode interelectrode capacitance. The wavelength of the system is substantially determined by the grid circuit, 3 and 4, while the Lecher system, 1 and 2, acts chiefly as a voltage transformer. At resonance, i.e., when the distance between bridge 6, and grid connector 3, measures one or more half waves, standing waves of frequencies up to 200 megacycles may readily be demonstrated by neon indicators, 7.

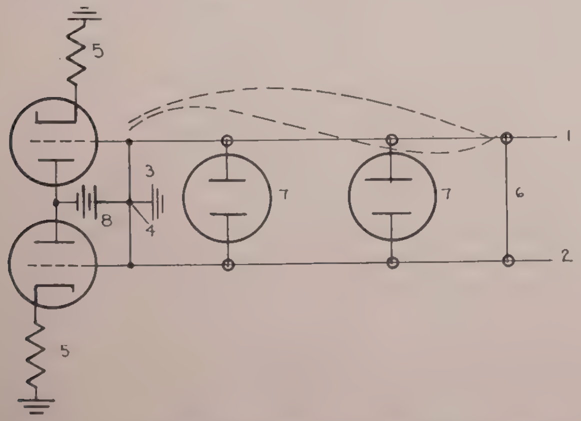

Fig. 13—An ultra-high-frequency cathode-follower oscillator.

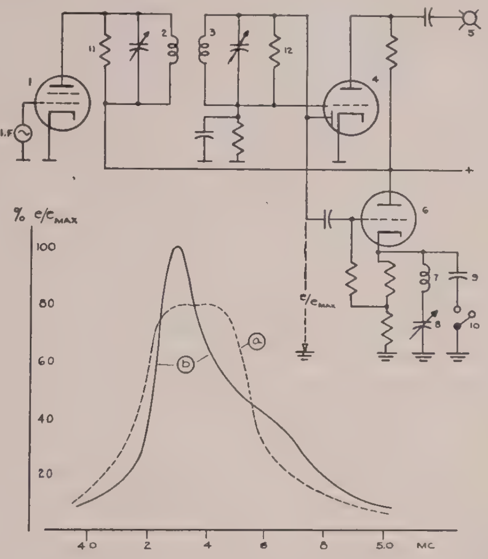

Fig. 14—Band-pass filter with cathode-follower termination.

Still higher frequencies (700 megacycles and above) have been obtained with acorn tubes as ocillating cathode followers. The grid circuit was a concentric tuning stub, and the cathode was loaded only by its own electrode capacitance. At very high frequencies the most serious limitation was found to be the inductance of the plate-to-ground connections and of the plate leads within the tube.[23]

C. Filter with Cathode-Follower Termination

It was shown in Section III, *A*, that a cathode follower with a series-resonant circuit in the cathode branch suppresses frequencies below and enhances frequencies above cathode resonance. This may be demonstrated in a circuit such as Fig. 14. This is an ordinary receiver circuit for amplitude modulation which has an intermediate-frequency output stage 1, a band-pass filter 2 and 3, and damping resistors 11 and 12, as well as a second detector 4, of the diode type. Audio frequency is thus available at 5 and the over-all characteristic would be symmetrical around center frequency as shown in curve *a* of Fig. 14, if tube 6 were omitted.

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[22] The grid-bias resistors 5, of 150 ohms are large as compared to the reactance of the cathode capacitance at these very high frequencies. This condition may be assured by the use of cathode chokes.

[23] Eduard Karplus, "Wide-range tuned circuits and oscillators for high frequencies," PROC. I.R.E., vol. 33, pp. 426–441; July, 1945.

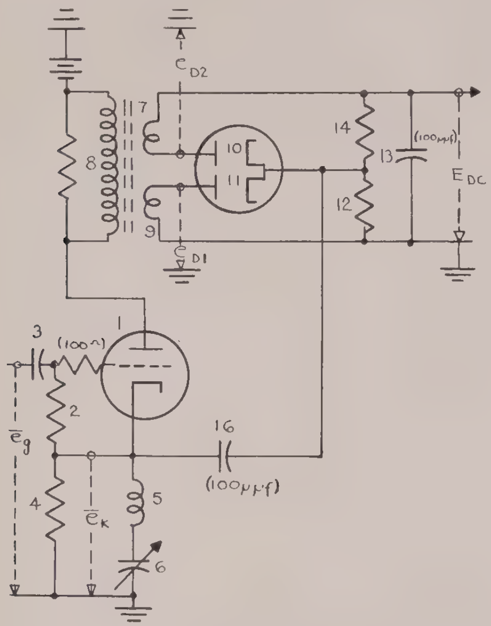

Fig. 15—Cathode-follower detector for frequency modulation.

The complex cathode follower 6 is connected in parallel to the filter output and its cathode circuit 7 and 8 is tuned to the center frequency. The symmetrical band-pass filter characteristic of Fig. 14*a*, is thereby transformed into a sloping filter characteristic of curve *b*, with the center frequency as the turning point. As a result, it becomes possible to receive frequency-modulated signals with detectors for amplitude modulation.

The circuit is reconverted into an amplitude-modulation receiver either by the switch 10, which connects a large by-pass capacitor 9 across the cathode circuit, or else by breaking the heater circuit of the cathode-follower stage. In both cases there is no change in the cathode-follower input capacitance and thus no detuning of the band-pass filter.

IV. A Cathode-Follower Detector for Frequency Modulation

Another application of the complex cathode follower for frequency modulation is shown in Fig. 15. It presents a frequency discriminator[24] which compares favorably with more conventional systems, as it combines detector and limiter action. The circuit is basically a complex cathode follower with tuned load 5 and 6, the output voltage of which is rectified through a double diode 10 and 11, in differential arrangement. Two equal and opposite exciter voltages e_{D1} and e_{D2} are induced in series with the diodes by an aperiodic radio-frequency plate transformer 7, 8, and 9. The audio output voltage across resistor *12–14* is then zero at cathode resonance and is proportional to frequency excursions below and above that frequency.

The operation of the system is illustrated in the form of a vector diagram in Fig. 16. **OA** is the input voltage to the cathode-follower grid and is assumed to be 1 volt. Cathode voltage and plate current may then be found for any frequency by plotting the product g_mX_k along a horizontal line through point *B*, 1 volt above the origin (**BD**$=g_mX_k$). The line **OD** then intersects at **E**, the circle having **OA** as diameter. This yields immediately the cathode voltage $\bar{e}_k=$**OE**. The plate current, $\bar{i}_p=$**OF**, is obtained by locating a point *F* diametrically opposite to **E**. The exciter voltages **OH** and **OG** appear at right angles to the plate current since they are induced in the plate transformer (7, 8, and 9, of Fig. 15), which operates far below resonance. It may be shown that the amplitudes of the diode voltages, e_{D1} and e_{D2}, have no influence on the output of the system so long as they exceed the input amplitudes; i.e., so long as points **G** and **H** lies at or beyond the circle with center **O** and radius **OA**. The form of the discriminator characteristic for 1-volt input may then be found by laying off twice the length **OE**, vertically through **D**, giving the ordinate **LK**. This yields a curve of the type[25]

$$y = 2x/\sqrt{1 + x^2} \tag{40}$$

which is linear and symmetrical with respect to center

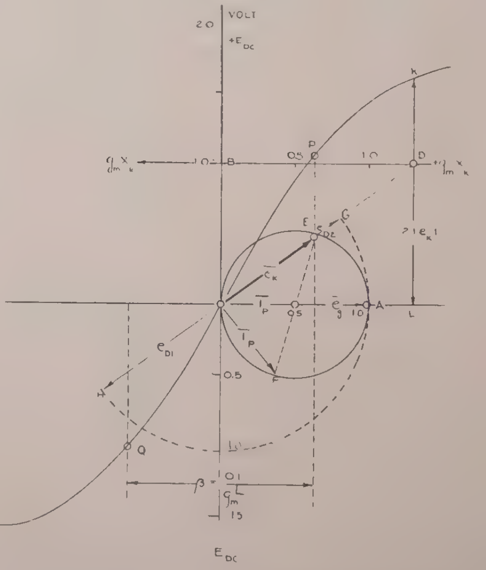

Fig. 16—Vector diagram for cathode-follower detector.

frequency over a range $P-Q$, where output and input amplitudes are equal. Above that value, output saturation begins, but there is no reversal of the characteristic as found in conventional discriminator circuits.

The limiter action of the system may be explained by the fact that the output depends on the cathode swing rather than on the exciter voltages for the diode plates.

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[24] United States Patent pending.

[25] See Appendix V, equation (64).

From the cathode-follower diagram of Fig. 2 it is apparent that the amplitudes between cathode and ground are confined between the points **O** and **N**. This presumes that the grid-leak resistor 2; is high enough to allow for additional grid bias for strong signals. In actual tests, constant output was obtained for input voltages ranging from less than one volt to several volts.

A disadvantage of the circuit of Fig. 15 is the excessive bandwidth necessary to cover fully the linear output range. For a simple two-reactance arm with a cathode inductance L_k, the bandwidth[26] is

$$b = 160/L_k \cdot g_m. \tag{41}$$

This is listed in Table III for various tubes and for a center frequency of 4.2 megacycles, to which is tuned a resonant circuit with $L_k = 71$ microhenrys, $C_k = 20$ micromicrofarads, and negligible series resistance.

TABLE III

CATHODE-FOLLOWER DETECTOR FOR FREQUENCY MODULATION

$f_c = 4.2$ megacycles	$L_k = 71 \mu H_y$	$C_k = 20$ micromicrofarads	$\rho_k = 0$
Tube Type	Transconductance g_m (milliamperes per volt)	Bandwidth b (kilocycles) eq. (41)	Type of Circuit
6Q7 6R7	1.2 1.9	1840 1180	Internal diode-electrodes
6J5 6AC7 6J4	3.0 8.0 12.0	750 280 188	Separate double-diode necessary

With ordinary diode-triodes, (type 6R7) with $g_m = 2$ milliamperes per volt, it is difficult to obtain better selectivity than about 1 megacycle. There are two possible ways of improving the selectivity of the discriminator and obtaining bandwidths of 300 kilocycles or less: either a tube with high transconductance, or a more complex network may be used. Tubes with transconductances of 8000 micromhos and more are listed in Table III with the resultant selectivities for a two-reactance cathode arm, as computed from equation (41). The bandwidth may be expanded by the use of resistor in series with the cathode arm. The other alternative uses a four-reactance network as shown in Fig. 17 as complex load for a standard triode. This circuit yields high selectivity and adjustable bandwidths down to 200 kilocycles in connection with tubes of moderate transconductance. Fig. 18 shows experimental data obtained with a 6J5 tube in the circuits of Fig. 15 and Fig. 17. Curve *a* of Fig. 18 shows the response of the simple detector circuit with only two reactance elements; its bandwidth is 800 kilocycles. In the system with four reactances, the bandwidth can be adjusted between 400 and 200 kilocycles by reducing the separation between the two antiresonant frequencies of the cathode network. (Curves *b* and *c* of Fig. 18.) In actual practice, the circuit of Fig. 15 may be more attractive because it has only one tuning element for adjustment to center frequency.

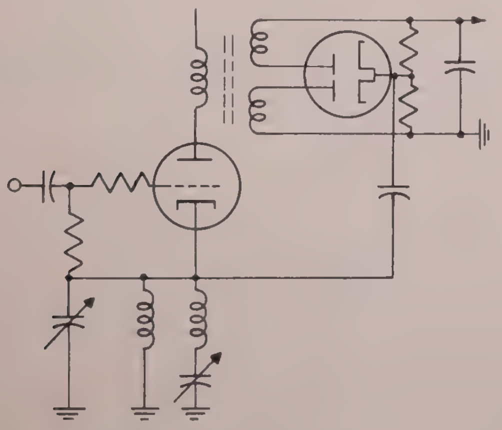

Fig. 17—Equivalent cathode-follower discriminator with four-reactance network.

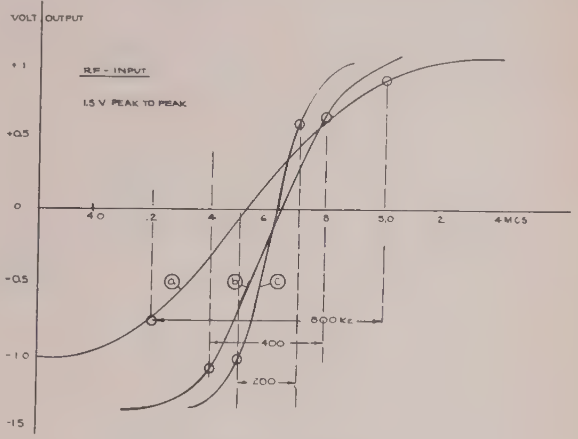

Fig. 18—Selectivity and bandwidth for the circuits of Figs. 15 and 17.

This frequency-modulation detector circuit is somewhat more involved than the conventional double-diode discriminator, but it has, in addition to limiter action, the advantage of higher input impedance and lower output impedance. The former permits higher gain from the output stage of the intermediate-frequency amplifier and the latter allows for the transmission of higher modulation frequencies. In addition to this, the cathode-follower circuit provides efficient filtering of the plate power supply, and thereby a beneficial reduction of background noise.

CONCLUSION

In this paper, cathode-follower circuits have been presented in various technical applications. The cathode-output amplifier is firmly established as impedance transformer and signal equalizer for wide-band transmission. It may also be used with output transformer in an audio amplifier to yield a gain of about 7:1 with good fidelity. As an interstage coupling element, the cathode follower is less attractive since equivalent gain may be more economically obtained by conventional coupling networks. A cascade of several cathode-follower tetrodes may be used to advantage for signal synthesis in transmitters for television and multiplex communication.

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[26] See Appendix IV, equation (57).

Next it is shown that the cathode follower with capacitive load becomes an oscillator at or above the cutoff frequency, where it ceases to be a true signal repeater. This cathode-follower oscillator lends itself to theoretical treatment which indicates the existence of an optimum design. Under such conditions, oscillations with good wave form may be obtained up to frequencies of 700 megacycles or more without external feedback.

With tuned circuits in the cathode arm, the cathode follower may be used as a detector for frequency-modulated signals. Some of these circuits compare quite favorably with more conventional discriminators for frequency modulation.

APPENDIX I

Expressions for transconductance S and source impedance R_i of the cathode follower are readily derived from the equivalent triode characteristic, equations (1) and (6), by partial differentiation.

$$S = \frac{\partial I_k}{\partial e_g} = \frac{dI_k}{dE_g}\cdot\frac{\partial E_g}{\partial e_g} = g_m[1 - SR_k(1 + 1/\mu)]\cdot \qquad (42)$$

This is solved for S

$$S = \frac{1}{(1 + 1/\mu)R_k + 1/g_m}\cdot \qquad (43)$$

The linearization of the original triode characteristic is reflected by the fact that $1/g_m$ becomes negligible for large cathode resistors R_k. The source impedance of the cathode follower is found in a similar manner

$$R_i = -1\Big/\frac{\partial I_k}{\partial e_k} \qquad (44)$$

$$\frac{\partial I_k}{\partial e_k} = \frac{dI_k}{dE_g}\cdot\frac{\partial E_g}{\partial e_k} = -g_m(1 + 1/\mu) \qquad (45)$$

$$R_i = \frac{1}{g_m(1 + 1/\mu)} < 1/g_m. \qquad (12)$$

The transconductance of the cathode follower may now be conveniently expressed in terms of source and load resistances

$$S = \frac{\mu}{\mu + 1}\cdot\frac{1}{R_i + R_k}\cdot \qquad (46)$$

APPENDIX II

CATHODE-FOLLOWER AMPLIFIER WITH OUTPUT TRANSFORMER

A cathode follower with a source impedance $R_i = 1/g_m$ feeds the primary inductance L_1 of the transformer with an input voltage which is down to 70.7 per cent of its maximum value at the low frequency F, where

$$2\pi FL_1 = R_i. \qquad (47)$$

At or above the highest frequency f, resonance occurs at the secondary

$$f \leqq 1/2\pi\sqrt{L_2C_2}. \qquad (48)$$

A damping resistor R_2 maintains uniform amplification

$$R_2 = 1/2\pi fC_2. \qquad (49)$$

The voltage gain by transformation $a = \sqrt{L_2/L_1}$ thus becomes

$$a \leqq \sqrt{g_m \cdot R_2 \cdot F/f}. \qquad (50)$$

This yields $a = 6{:}1$ when $F = 120$, $f = 10{,}000$, $g_m = 10{,}000$ micromhos, $C_2 = 50$ micromicrofarads, $R_2 = 0.3$ megohm. Frequencies below F are emphasized through resonance with the input coupling capacitor, such that the actual band extends to frequencies lower than F.

APPENDIX III

COMPLEX CATHODE IMPEDANCE

The capacitor C_{gp}, (Fig. 1), is considered as part of the outer circuit. The input admittance through the grid-cathode capacitance C_g then becomes

$$1/\bar{z}_i = \bar{\imath}_g/\bar{e}_g = pC_g\frac{\bar{e}_g - \bar{e}_k}{\bar{e}_g}$$

$$= pC_g/1 + \bar{z}_k(pC_g + g_m);\ p = j\omega. \qquad (51)$$

A complex cathode impedance is assumed (circuit c of Fig. 9).

$$\bar{z}_k = \rho_k + jX_k. \qquad (52)$$

Equation (52) may then be rationalized as follows:

$$1/\bar{z}_i = 1/R_i + pC_i \qquad (53)$$

with the components

$$R_i = \frac{1}{\omega C_g}\cdot\frac{(1 + \rho_k g_m - \omega C_g X_k)^2 + (X_k g_m + \omega C_g \rho_k)^2}{X_k g_m + \omega C_g \rho_k} \qquad (54)$$

$$C_i = C_g\frac{1 + \rho_k g_m - \omega C_g X_k}{(1 + \rho_k g_m - \omega C_g X_k)^2 + (X_k g_m + \omega C_g \rho_k)^2}\cdot \qquad (55)$$

These components of the input admittance are represented in Fig. 11 under the assumptions that

$$\rho_k = 0, \text{ and } \omega C_g < g_m. \qquad (56)$$

The two critical frequencies f_1 and f_2 are solutions of the equations

$$X_k = +\frac{1}{g_m}\cdot\frac{1 + \rho_k g_m(1 - \omega C_g/g_m)}{1 + \omega C_g/g_m} \qquad (33a)$$

and

$$X_k = -\frac{1}{g_m}\cdot\frac{1 + \rho_k g_m(1 + \omega C_g/g_m)}{1 - \omega C_g/g_m}\cdot \qquad (33b)$$

At these frequencies, the input resistance R_i (equation (54)), assumes a minimum value of

$$R_{i\min} = \pm 2\left[\frac{1 + \rho_k g_m}{\omega C_g} - X_k\right] \qquad (34a)$$

which is positive or negative along with the sign of X_k at f_1 and f_2, respectively.

APPENDIX IV

FILTER WITH CATHODE-FOLLOWER TERMINATION

For a tuned-cathode circuit with L_k and C_k in series, (circuit *c* of Fig. 9) the cathode reactance is

$$X_k = \sqrt{\frac{L_k}{C_k}}\,(x - 1/x) \qquad (32)$$

where

$$x \doteq f/f_{\rm res} \quad \text{and} \quad f_{\rm res} = 1/2\pi\sqrt{L_kC_k}.$$

The bandwidth $b = f_2 - f_1$ of the cathode-follower filter is found by combining equations (33) and (34), assuming $b/f_{\rm res} < 1$.

$$b = (1 + g_m\rho_k)/2\pi L_k g_m. \qquad (57)$$

The bandwidth is found to be independent of the tuning capacitance C_k. It is tabulated in Table III for negligible resistance in the cathode impedance. Apparently an adjustable resistance ρ_k, in series with the cathode reactances, may be used to stretch this band arbitrarily.

APPENDIX V

A CATHODE-FOLLOWER DETECTOR FOR FREQUENCY MODULATION

In Fig. 15, cathode voltage and plate current are

$$\bar{e}_k = \bar{e}_g \frac{jg_mX_k}{1 + jg_mX_k} \qquad (58)$$

$$\bar{\imath}_p = \bar{e}_g \frac{1}{1 + jg_mX_k}. \qquad (59)$$

The diodes receive exciter voltages

$$\bar{e}_{D1} = j\omega M\bar{\imath}_p = \bar{e}_g \frac{j\omega M g_m}{1 + jg_mX_k} \qquad (60)$$

$$\bar{e}_{D2} = -\bar{e}_{D1} \qquad (61)$$

from the split secondary of the plate transformer. (Mutual inductance M.) These are in phase with $\bar{e}_k$ and $-\bar{e}_k$, respectively (see vector diagram, Fig. 16). The rectified currents are proportional to the absolute value of the diode-to-cathode voltages

$$I_1 = 2/R\cdot|\bar{e}_{D1} - \bar{e}_k| = \frac{2|e_g|}{R}\cdot\frac{g_m[\omega M - X_k]}{\sqrt{1 + (g_mX_k)^2}} \qquad (62)$$

$$I_2 = 2/R\cdot|\bar{e}_{D2} - \bar{e}_k| = \frac{2|e_g|}{R}\cdot\frac{g_m|\omega M + X_k|}{\sqrt{1 + (g_mX_k)^2}}. \qquad (63)$$

The total direct voltage across the common output resistor R is proportional to the difference of these two current contributions

$$E = \frac{R}{2}(I_2 - I_1) = 2|e_g|\cdot\frac{g_mX_k}{\sqrt{1 + (g_mX_k)^2}}. \qquad (64)$$

This is the analytical expression for the detector characteristic (equation (40)), as it is plotted in the vector diagram of Fig. 16, and verified by the measurements of Fig. 18.

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Television Resolution as a Function of Line Structures[*]

MADISON CAWEIN[†], SENIOR MEMBER, I.R.E.

Summary—A labor-saving method is developed for obtaining complex wave shapes by means of evaluating integrals. This method is applied to the problem of square-wave pulse shapes containing large numbers (not infinite) of harmonic components. The method is applicable for any value of cutoff frequency, expressing this value in terms of electrical degrees. A theory of "aperture processes" is developed by mathematical induction, and it is shown that the finite width of a pulse (line structure of an image) is mathematically equivalent to the pulse width obtained from an image of infinite detail scanned by a finite aperture. The finiteness of the pulse width in either case, constitutes in itself an aperture process. An approximate rule of thumb is given for defining the resolution of an image, in terms of the lines-of-resolution involved in each process imposed upon it during optical-electrical conversions.

This is applicable to discussions concerning the effect of quality in the televised subject as it affects quality in the received image: as, for example, concerning the relative merits of 16- and 35-millimeter film.

I. INTRODUCTION

MANY television problems may be solved by utilization of pulse theory. One of the more pertinent problems at the present time concerns the degradation of an image, having a certain original resolution, after transmission through various optical and electrical systems. The approach is purely theoretical. The results explain *qualitatively* why there is a difference in resolution between television images transmitted from 35- or 16-millimeter film, although the resolution capability of either film is considered to be superior to television resolution in the 4.50-megacycle channels assigned at present. No quantitative data are as yet available to check the results expressed here by a

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[†] Farnsworth Television and Radio Corporation, Fort Wayne, Indiana.

rule-of-thumb formula which is developed in the last section of the paper. In spite of the lack of quantitative experimental confirmation, it was thought that this subject would be of general interest in television circles.

Fig. 1 shows two cycles of a train of waves which are of particular interest in television problems. This wave has been called variously by many names, among which the more prominent are pulse, blanking-signal, and pedestal. Although the wave diagram of Fig. 1 is a mathematically ideal representation, electric waves of potential, which are closely approximate to this ideal may be generated.[1–7] It is the purpose of this paper to present some useful computations in regard to Fig. 1, and concerning image resolution.

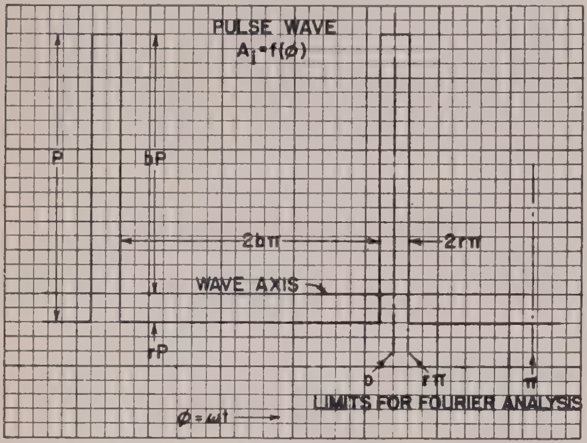

Fig. 1—Graph of a finite rectangular pulse.

The pulse wave has three distinguishing characteristics: amplitude, duration, and frequency of repetition. It will be noted that the positive and negative excursions of the wave are different in amplitude, and also that the durations of the positive and negative portions are different.[8] The ratio of these durations is inversely proportional to their amplitudes because of the equality of areas. The ratio of the time-of-least-duration to the time-of-the-entire-cycle will be defined by the symbol r (the fraction of the cycle for which the positive peak endures, in Fig. 1). The ratio of the time-of-greatest-duration to the period of the cycle will be defined by the symbol b. Thus, $b+r=1$. The "r index" is a "duration index" of the wave, and waves will be described by reference to this r index, where r is a fraction always less than unity.

The equation of the pulse wave may be obtained by the methods of Fourier analysis.[9,10]

The equation of an infinite train of pulse waves, in the steady state, at any instant of time, is (see Appendix A for derivation)

$$A_i = \overset{\text{(I)}}{2rP}\sum_{n=1}^{\infty}\overset{\text{(II)}}{\frac{\text{Sin } nr\pi}{nr\pi}}\cdot\overset{\text{(III)}}{\text{Cos } n\omega t}. \tag{1}$$

It will be observed that there are three factors composing this equation. In the order of their occurrence in (1), these factors are:

(1) The constant term, the magnitude of which has been properly chosen so that the sum of the infinite number of harmonic components will always add (at the proper phase point) to give the proper peak value to the wave.

(2) The harmonic-amplitude-adjusting coefficient, which is independent of the time, but is a transcendental function of nr.

(3) The functional component, which is dependent upon the time.

II. Amplitude of Harmonic Components

A pulse wave defined by (1), having a duration index r, will be referred to as an r-type wave. The second factor, Sin $nr\pi/nr\pi$, under column II of (1), has the property of modulating the amplitudes of the harmonic components. It will be noticed that this factor is of a peculiar form. It is zero at every value of n which is an integral multiple of $1/r$, since $nr\pi$ is then an integral multiple of π. These zero points are called the elided frequencies of the wave. They represent points at which the amplitudes of the harmonic components fall to zero. The similarity to discussions on the effect of finite aperture widths will be apparent.[11] In fact, the width of the pulse and the width of an aperture are identical as mathematical concepts.

The first, second, and third zero frequencies of an r-type wave occur at $n=1/r$, $2/r$, $3/r$. Because of the fact that the harmonic amplitudes are positive in the angular interval up to the first zero, and negative in the interval from the first zero to the second zero, then if only those components up to the first zero frequency are present, the resultant wave form will overshoot the actual wave form which it simulates (see curve a of Fig. 2).

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[6] "Television bibliography," *Electronics*, p. 265; August, 1932.

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[8] Mathematically, a wave motion is a cyclical disturbance-with-time, in some form of space (energy space, momentum space, etc.). This occurs in such fashion that there is no net change in displacement from cycle to cycle. The locus of zero displacement is the axis of the wave. This divides the wave into positive and negative portions in which the time-integral-of-displacement in each cycle is zero; i.e., the positive and negative areas are equal. Electrically, the displacement of charge is involved, and the rule of equality-of-areas may be interpreted to mean that there is no net transfer of charge, in an alternating circuit, from one cycle to another.

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In Fig. 3 there has been plotted curve 1, which represents the amplitudes of the harmonic components of the pulse wave of (1) for the first two angular intervals. The product $nr\pi$ is a pure number, and may be represented as an angle: $nr\pi=\phi$ radians. The function

$$S_1 = \operatorname{Sin} \phi/\phi \tag{2}$$

is a discontinuous function when ϕ is defined in this manner.[12] Graph 1 of Fig. 3 is a plot of the envelope of the discontinuous function S_1.

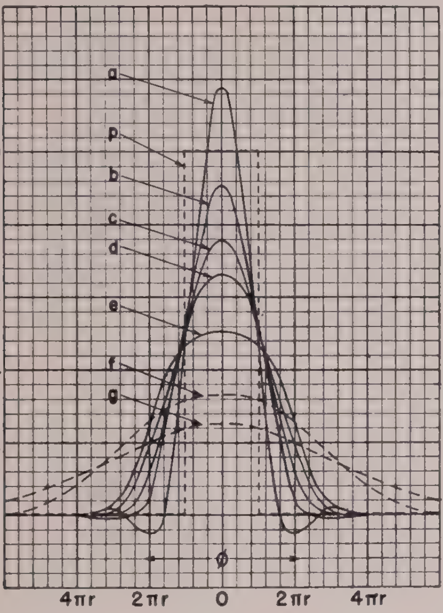

Fig. 2—Finite pulse shapes. Synthesis of harmonics of a finite pulse, with cutoff at 1st elided frequency.

p = rectangular pulse for infinite number of harmonics (neglecting Gibb's phenomenon). Electrical signal will have this pulse form, if a finite illuminated area of width index r is scanned by a photocell having an infinitesimal aperture. This is the graph of (1) for $N=\infty$.
a = shape of the pulse P for a finite number of harmonics. This is the graph of (1) for $N=1/r$.
b = electrical signal will have this pulse form if a finite, illuminated area is scanned by a photocell having a finite aperture of the same width, and if electrical cutoff is at first elided frequency. This is the pulse shape from a picture element scanned by one aperture of element width.
c = pulse shape from a picture element scanned by two apertures in tandem, each of element width.
d = pulse shape from a picture element scanned by three apertures in tandem, each of element width.
e = pulse shape from a picture element scanned by one aperture of twice element width.
f = pulse shape from a picture element scanned by one aperture of three times element width.
g = pulse shape from a picture element scanned by one aperture of four times element width.

The function S_1 approaches a continuous function as a limit, when r, the index of pulse width, approaches zero. The integral of S_1, as a continuous function, is known as the sine integral.[13] Fig. 3 (graph 1) is a graph of the integrand of the sine integral, which is the envelope of the amplitudes of the n harmonic components of a finite, pulse wave. It will be noted that the amplitude of S_1 goes through zero many times, and reverses sign as it does so, and also that the successive maxima of S_1 diminish steadily as ϕ increases, ϕ being proportional to n.

For each width of the pulse; i.e., for each value of r there is only a finite number of harmonics in each interval of S_1. The number of these harmonics is determined by the equation $nr=1$. As the duration index r approaches zero; that is, as the duration of the pulses becomes very short, the parameter $\phi=nr\pi$ may be considered as a continuous variable. Actually, ϕ is a "quantized" variable which changes its value in discrete jumps, each of magnitude $\Delta\phi=r\pi$. The value of ϕ at any harmonic frequency will be determined by the number of these jumps, which number is equal to n, the order of the harmonic.

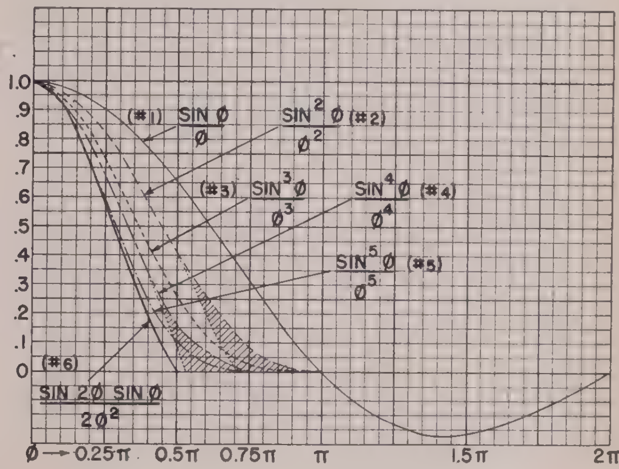

Fig. 3—Graphs of the function S_1 and related functions describing the harmonic contents of pulses subjected to successive aperture processes.

1 = relation of harmonics in a rectangular pulse, up to the second zero frequency. (One aperture process.)
2 to 5 = relation of harmonics, up to the first zero frequency, for 2, 3, 4, and 5 in tandem aperture processes.
6 = approximation for S_1 when two aperture processes are involved, having a ratio of 1:2.

The areas of the sine integral have been tabulated for many values of $\phi=nr\pi$ and may be interpolated from tables for other values.[14,15] The abscissa scale to which Fig. 3 has been plotted is such that zero frequencies occur whenever the values of nr are integral; i.e., whenever ϕ is a multiple of π.

The units of abscissa may be converted rapidly to frequency, in case the periodicity of the wave and its duration index are known. For example, if it is desired to know the first zero frequency of a 60-cycle pulse wave of 0.1 duration ($r=0.1$), choose $n=1/r=10$ such that the first zero frequency occurs at the 10th harmonic of the fundamental; i.e., at 600 cycles. The harmonic amplitude of any component can be found immediately for the 10 values of ϕ, at which harmonics exist, by dividing each interval into 10 equal parts. The amplitudes of the harmonics are read from the curve at these points. It will be appreciated that, since harmonics are always integral multiples of the fundamental, there are only discrete points along the curve within each interval which are useful for any particular value of r. The reciprocal of the value of r for a pulse wave denotes the number of points from zero to the first zero-frequency and between zero frequencies, where harmonics actually exist. The units of abscissa in the diagram of Fig. 3 are units of $r\pi$, for any particular wave. The smaller the value of $r\pi$, the more points along the curve are needed to describe its harmonic content.

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[12] According to this definition, ϕ is not a continuous variable but has only a discrete number of values, depending upon the number of values which n can assume (up to $n=1/r$), in each interval.

[13] Ernst A. Guillemin, "Communication Networks," vol. 2, John Wiley and Sons, Inc., New York, N. Y., 1935, p. 478.

[14] Federal Works Agency, Work Projects Administration for the City of New York, "Tables of Sine, Cosine and Exponential Integrals," vol. 2, 1940.

[15] See also N. R. Jørgensen, "Undersøgelser over Frequensflader og Korrelation," A. Busck, København, 1916, pp. 157–175, for table of $1/\pi\ Si\ (nx)$.

In Fig. 2*a* the synthesis of the pulse wave is shown for a finite sum of harmonics, added up to the first zero frequency. "Plotting" cutoff is, in this case, at $1/r$ harmonics. The curve holds approximately for any value of r smaller than 0.1, i.e., for narrow pulse width relative to the spacing between the pulses. The curve becomes more exact, the smaller the value of r. Thus Fig. 2*a* is approximately a plot of a single cycle of an 0.1 pulse wave, which would be obtained by adding up all the components to the 10th harmonic, at every phase point, and neglecting all other components. The pulse wave of Fig. 2*a* represents the output of an imaginary electrical system with flat transmission and linear phase characteristic up to the first zero frequency, whether this frequency be at the 10th or 10,000th harmonic frequency, when the input wave is of the form shown in Fig. 1, (or Fig. 2*p*). Since neither the input wave nor the filter can exist, the pulse shape of Fig. 2*a* is confined to theoretical interest and has been developed as an aid to the discussion to be undertaken in Section III.

III. SYNTHESIS OF WAVES FROM THEIR HARMONIC COMPONENTS

The factor of the form $\sin\phi/\phi$ is a function of $\phi = nr\pi$. When r is small, $r\pi$ may be treated as an increment. Let $r\pi = \Delta\phi$. Multiplying (1) by unity $= \Delta\phi/\Delta\phi = \Delta\phi/r\pi$.

$$A_i = \frac{2P}{\pi}\sum_{n=1}^{\infty}\frac{\sin\phi}{\phi}\cdot\Delta\phi \cos n\omega t. \tag{3}$$

This equation represents the pulse wave for infinite bandwidth. Such form of the equation suggests that the amplitude of the wave could be found for any finite number of harmonics by expressing ωt in terms of ϕ, and integrating, or adding areas under the summation sign, up to a finite number of harmonics ($n = N$). This is a substitute for the laborious process of actually adding N harmonic components, where N is a very large number.

Since ϕ is proportional to n, the limits must be changed from units of n to units of ϕ; i.e., from $n = 1$ and $n = N$, to $\phi = 0$ and $\phi = \phi_1$. The summation, then, represented in (3) may be replaced by an integral. ($\Delta\phi$ may be replaced by $d\phi$ when r is small.)

$$A_i \text{ (finite cutoff)} = \frac{2P}{\pi}\int_0^{\phi_1}\frac{\sin\phi \cos n\omega t\, d\phi}{\phi} \tag{4}$$

where $\phi_1 = Nr\pi$.

Equation (4) is not in condition for simple integration. It will be necessary to determine ωt in terms of $r\pi$ and to set the limits of integration, according to the number of harmonic terms it is desired to add. There are only a few values of ωt for which it is necessary that the amplitude of the wave be determined, in order to plot its general shape. It will be sufficient to plot the curve at six points: (a) $\omega t = 0$; (b) $\omega t = r\pi/2$; (c) $\omega t = r\pi$; (d) $\omega t = 3r\pi/2$; (e) $\omega t = 2r\pi$; and (f) $\omega t = 3r\pi$. (Calculations for curves of Fig. 2 are shown in Appendix B.)

This number of points is believed to be sufficient to indicate the shape of a universal, synthetic pulse wave, having any large number of harmonic components up to the 1st elided frequency. Integration limits are determined by the cutoff frequency (at the Nth harmonic). In substituting the values for ωt in (4), six integrals are obtained.

$$A_a = \frac{2P}{\pi}\int_0^{\pi}\frac{\sin\phi}{\phi}\,d\phi \tag{5a}$$

$$\begin{aligned} A_b &= \frac{2P}{\pi}\int_0^{\pi}\frac{\sin\phi}{\phi}\cdot\cos\phi/2\,d\phi \\ &= \frac{4P}{\pi}\int_0^{\pi/2}\left(\frac{\sin\phi}{\phi} - \frac{\sin^3\phi}{\phi}\right)d\phi \end{aligned} \tag{5b}$$

$$A_c = \frac{2P}{\pi}\int_0^{\pi}\frac{\sin\phi\cos\phi}{\phi}\,d\phi = \frac{P}{\pi}\int_0^{2\pi}\frac{\sin\phi}{\phi}\,d\phi \tag{5c}$$

$$\begin{aligned} A_d &= \frac{2P}{\pi}\int_0^{\pi}\frac{\sin\phi\cos 3\phi/2}{\phi}\,d\phi \\ &= \frac{4P}{\pi}\int_0^{\pi/2}\left(\frac{\sin\phi}{\phi} - \frac{5\sin^3\phi}{\phi} + \frac{4\sin^5\phi}{\phi}\right)d\phi \end{aligned} \tag{5d}$$

$$\begin{aligned} A_e &= \frac{2P}{\pi}\int_0^{\pi}\frac{\sin\phi\cos 2\phi}{\phi}\,d\phi \\ &= \frac{2P}{\pi}\int_0^{\pi}\left(\frac{\sin\phi}{\phi} - \frac{2\sin^3\phi}{\phi}\right)d\phi \end{aligned} \tag{5e}$$

$$\begin{aligned} A_f &= \frac{2P}{\pi}\int_0^{\pi}\frac{\sin\phi\cos 3\phi}{\phi}\,d\phi \\ &= \frac{P}{\pi}\left[\int_0^{4\pi}\frac{\sin\phi}{\phi}\,d\phi - \int_0^{2\pi}\frac{\sin\phi}{\phi}\,d\phi\right]. \end{aligned} \tag{5f}$$

If $N = 1/r$, $\phi_1 = \pi$; and the values calculated from (5) correspond to those which would be obtained by adding N harmonics. If $N = 2/r$, $\phi_1 = 2\pi$. Using $\phi_1 = 2\pi$ in (4) corresponds to adding harmonic components up to the second elided frequency. Calculations for $\phi_1 = \pi$, in (5), are shown in Appendix B, and plotted in Fig. 2 as pulse *a*.

As an example of the error which is introduced by using an integral method for values of r as large as 0.1, the negative overshoot for the universal wave of Fig. 2*a*' at the points $\pm r\pi$, is 5 per cent of the total amplitude of the wave when computed by (5); whereas, the actual addition of the ten Fourier components at this point would give a 6 per cent negative overshoot.

For the theoretical case where $r=0$ (of academic interest only) the equation of the infinitesimal pulse may be obtained from (1) by letting r approach zero, in which case the Sin ϕ/ϕ factor becomes unity. The shape of the *infinitesimal* pulse, obtained by adding a *finite* number of harmonic components $N=1/r$ is shown in Fig. 4 (see Appendix D).

IV. Pulse Signals Through Apertures

A television system is an optical system for the transmission of images through moving apertures. The finite size of the aperture causes degradation of the image, as is well known.[11]

An analysis of the shape of an electrical pulse, from an illuminated point in the image on a television pickup tube, is of interest in regard to aperture distortion. Since a true mathematical point has no reality, that is, no shape and no dimensions, it is best to define the shape and dimensions of the points of light in the image to be discussed, and thus avoid a philosophical impasse. The shape of a point (a picture element) will be defined as "square." If an image is divided into parallel lines or strips and these lines are strung out end to end, the light intensities of picture elements across the strip image have an order of spacial distribution, which may be converted to an order of time distribution, by means of the repetitive scanning process. This implies that observation of the image is made through a small, square window which moves uniformly along the line. The uniform motion along the line corresponds to scanning speed, and the small window corresponds to the scanning aperture.

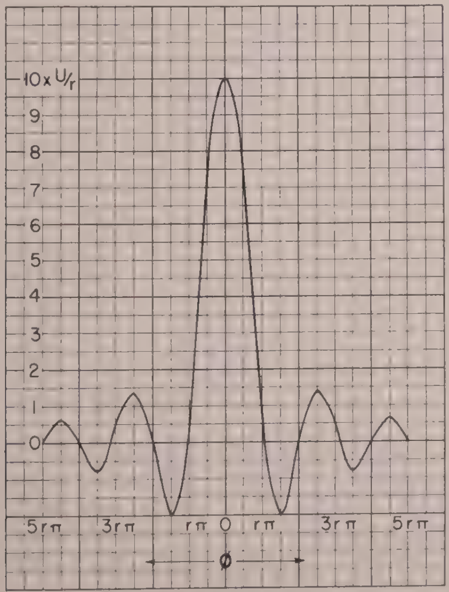

Fig. 4—Infinitesimal pulse for finite cutoff. Finite width of an infinitesimal pulse after transmission through a filter with sharp cutoff at a harmonic defined by $N=1/r$.

Consider the image to be made up of points of light and consider a device (an infinitesimal window) which would transform these points of light into instantaneous electrical potentials, as it traverses them. Light radiation will "pulse" through this window, at the frequency of scanning repetition. For the purpose of the analysis, it will be sufficient to consider only a single point of light in the entire picture: all other points could be considered on the same basis, having different relative phase and amplitude. The pulses of light have the wave form of Fig. 1.

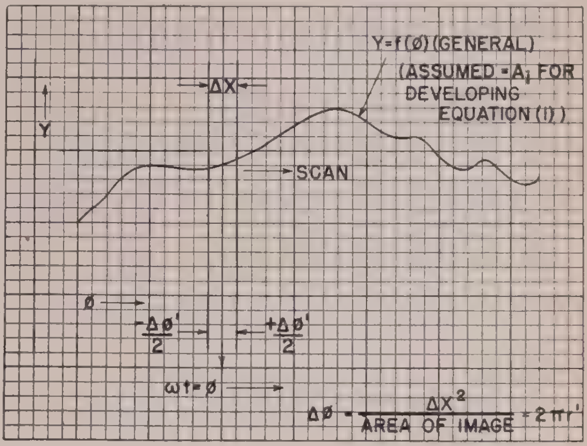
Y=f(Ø)(GENERAL)
(ASSUMED = Ai FOR DEVELOPING EQUATION (I))
ΔX
Y
SCAN
Ø
-ΔØ'/2
+ΔØ'/2
ωt=Ø
ΔØ = ΔX²/AREA OF IMAGE = 2πr'

Fig. 5—Graph of Y, which is intensity of light along scanning line, in its relation to the width of aperture, ΔX. ($Y=A_i$ for developing (1) of text.)

Thus, (1) describes the electrical components (in a photocell circuit) which come into being from each square point of light, during the scanning process (assuming an infinitesimal width of aperture). A finite scanning aperture has the effect of integrating (averaging) these components across its dimensions, in the direction of the scanning. Thus, to obtain the electrical output from a scanning system with a finite aperture, it is necessary to integrate (1) along the aperture in the direction of scanning. Let the dimensions of the aperture be $\Delta\phi' \times \Delta\phi'$ (Fig. 5). Then the integration must be performed between the limits $(\phi-\Delta\phi'/2)$, and $(\phi+\Delta\phi'/2)$. Let E_a = the electrical signal at field-scansion frequency, from the photo surface behind the finite aperture.

$$E_a = K' \int_{(\phi-\Delta\phi'/2)}^{(\phi+\Delta\phi'/2)} A_i' \cdot d\phi \quad (K' \text{ is a constant of the photocell}). \tag{6}$$

Integration and simplification of (6) yields (see Appendix C)

$$E_a = A' \sum_{n=1}^{\infty} \frac{\text{Sin } nr\pi}{nr\pi} \frac{\text{Sin } nr'\pi}{nr'\pi} \text{Cos } n\omega t. \tag{7}$$

Examination of (7) reveals that the only effect of a finite aperture is to introduce another factor of the form Sin ϕ'/ϕ' into the original harmonic composition of the pulse. The index of the aperture is r', the ratio of $\Delta\phi'$

to the total angular length (2π) of field-scanning-path across the entire strip image. The finite aperture, then, modifies the frequency spectrum which would be obtained through an infinitesimal aperture, during the scanning act, in the same manner that the *finiteness* of a pulse modifies the frequency spectrum of an infinitesimal pulse.

When the aperture index has the same value as the r index of the electrical pulses generated from image picture elements; that is, when the spot size is comparable to the picture element size which it is desired to reproduce, the modulating factor of the harmonic amplitudes in the frequency spectrum has a $\text{Sin}^2\,\phi/\phi^2$ form, as shown in curve 2 of Fig. 3. This is obvious from examination of (7) (substitute r for r'). It will be noticed that the first zero frequency of the aperture signal is identical to that of a finite picture element having the same size as the aperture (curve 1, Fig. 3). Curves 1 and 2 of Fig. 3 do not differ greatly from one another; only the higher-frequency components of the electrical spectrum have been modified by the finite aperture. For this reason, it is frequently stated that the quality of television pictures can be improved by aperture compensations; i.e., a rising frequency characteristic in the ratio of curve 1 to curve 2 in Fig. 3, *up to the first zero frequency*.

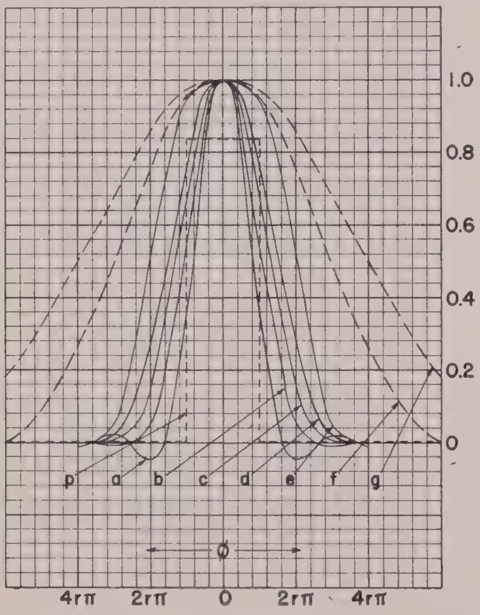

Fig. 6—Pulse shapes of Fig. 2. Equalized in amplitude.

The pulse shape shown in Fig. 2 as curve a is an idealized shape, for two reasons: the first of these is that it represents the output of an infinitesimal aperture, which cannot exist in practice but can be approximated; and the second is that it represents this output up to only the first zero frequency of the pulse, which implies that a sharp cutoff filter having perfectly linear phase shift up to cutoff has been inserted in the electrical circuit. This latter device does not exist, but can be approximated. The closer these two ideal conditions are realized, the more nearly the output wave form will be as shown in curve a, Fig. 2.

If the aperture has the same width as the structure of the image, that is, the width of the dotted rectangle shown as $2\pi r$ in Fig. 2, then the harmonic content is described by curve 2 of Fig. 3 and the shape of the electrical output will be as shown in curve b of Fig. 2. This shape is computed by methods similar to those used in (5) (see Appendix B), substituting $\text{Sin}^2\,\phi/\phi^2$ for $\text{Sin}\,\phi/\phi$. If, in addition, the filter has a cutoff characteristic $\text{Sin}\,\phi/\phi$, then the harmonic content is described by curve 3 of Fig. 3, and the output wave shape will be as shown in curve c of Fig. 2, which is computed according to the theory by substituting $\text{Sin}^3\,\phi/\phi^3$ for $\text{Sin}\,\phi/\phi$ in (5).

Should the scanning aperture have a width of twice that shown in Fig. 2; that is, a width of $4\pi r$, the output for a sharp cutoff filter (no frequency limitation below the first zero frequency) would be as shown in curve e of Fig. 2 (harmonic content as in curve 6 of Fig. 3). All the curves of Fig. 2 have one common characteristic: the area of the original, dotted rectangle is conserved. This may be proved by integration (planimeter), or by theoretical reasoning beyond the scope of this paper. Should all these curves be equalized in amplitude by insertion of electrical amplifiers, they would appear as shown in Fig. 6, which gives a basis for calculating the widening effect of successive aperture processes. The effective width of a pulse will be defined in the usual manner, as the width of an equivalent rectangle. Since the areas of the curves in Fig. 2 were equal (before equalizing the amplitudes), the areas of the curves in Fig. 6 must be inversely proportional to the amplitudes of the curves of Fig. 2, which means that the effective widths of the pulses are inversely proportional to their amplitudes, as calculated from (5).

V. SIGNIFICANCE, AS APPLIED TO RESOLUTION OF IMAGES[16–18]

Manufacturers rate film at 45 lines per millimeter, for fast film, or 70 lines per millimeter for slow, high-definition film. A 16-millimeter frame is 8 millimeters in height, and a 35-millimeter frame is 16 millimeters in height; thus, the manufacturer's rating of resolution for fast 16-millimeter film would be $8\times 45=360$ lines (720 television lines). In television practice, a line is defined as a scanning line, either black or white; whereas, it seems to be representative of optical practice to define resolution in terms of *distinguishable* lines; i.e., as black-and-white pairs. Thus, some film manufacturers define the resolution as one half of the value used in television.

If the quality of an image is defined by defining the number of "television lines" of resolution n_i, (such as

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[16] E. W. Engstrom, "A study of television image characteristics," Part I, PROC. I.R.E., vol. 21, pp. 1631–1652; December, 1933; Part II, vol. 23, pp. 295–311; April, 1935.

[17] A. V. Bedford, "Figure of merit for television performance," *RCA Rev.*; July, 1938.

[18] H. A. Wheeler and A. V. Loughren, "The fine structure of television images," PROC. I.R.E., vol. 26, pp. 540–576; May, 1938.

$n_i=800$, which is fairly representative of 16-millimeter film) and if the image has an effective[19] aspect ratio of A, then there are $A(n_i)^2$ picture elements. Each of these elements has, for scanning purposes, an electrical width (an r index) of

$$r = 1/A(n_i)^2. \tag{8}$$

(This value of r is the ratio of duration between one picture element in the picture, and the period of reproduction of a frame.)

The first, elided frequency of the electrical image, viewed through an infinitesimal aperture, is at the Nth harmonic of the frame frequency. If it were possible to produce electrical transmission filters with sharp cutoff at this frequency, the act of scanning would produce an electrical signal of the form shown in Fig. 2*a*.

If, however, the scanning aperture is square and of picture-element width (electrical index equal to r), then both the shape and the width of the pulse reproduced through the aperture will be modified from the shape of Fig. 2*a*. Mathematically, the modification could be plotted by performing the integrations in (5), substituting[20] the $\text{Sin}^2\,\phi/\phi^2$ curve for $\text{Sin}\,\phi/\phi$. This involves a great deal of work which can be avoided by another method of approximations.

It can be shown from Fig. 3 that the $\text{Sin}^2\,\phi/\phi^2$ curve, if extended, as in dot, to interesect the axis at 0.72π, would have a shape closely approximate to $\text{Sin}\,\phi/\phi$, with the exception of the cross-hatch area. The scale of plotting in electrical degrees is, however, contracted in the ratio 0.72/1, which corresponds to a decrease in the value of N, or to an increase in the value of r, to a new value $r''=r/0.72 \doteqdot \sqrt{2}\cdot r$.

The case in which two tandem scansions of a finite image are made by apertures which are finite and of width r indicates that the harmonic content is modified by a $\text{Sin}^3\,\phi/\phi^3$ curve (curve 3), which is approximately a $\text{Sin}\,\phi/\phi$ curve intersecting the axis at $0.62\,\pi$. This corresponds to an over-all width index of $r''' \doteqdot \sqrt{3}\cdot r$. For two apertures of width r, a filter[21] of width r, and an image of structure r, the curve is $\text{Sin}^4\,\phi/\phi^4$, and the approximate equivalent (four-in-tandem) aperture is $r'''' \doteqdot \sqrt{4}\cdot r$.

It is submitted that the finite structure of the original picture to be transmitted by electrical means imposes a restriction on the frequency band arising from the scanning process, which restriction is equivalent to that imposed by the scanning of a picture having infinite detail (if such could exist) by a finite aperture. This restriction will be defined as an "aperture process." Thus, to be explicit, *the finiteness of the structure of the picture itself in tandem with the finiteness of the aperture which scans it, is equivalent to the scansion of a perfect image by two apertures in tandem.*

By the method of mathematical induction, a rule of thumb may be evolved as to the "widening" effect, on a picture element, in subjecting the picture to any number of aperture processes. The transmission of frequency components up to the first elided frequency only is itself an aperture process, as stated in the preceding paragraph. Observation through an aperture of twice the width of picture element modifies the harmonic-adjusting coefficient of the frequency spectrum of a finite picture element as in curve 6 of Fig. 3 (pulse shape is plotted in Fig. 2*e*)

$$\text{Sin}\,\phi/\phi\ \text{Sin}\,2\phi/2\phi = \text{Sin}^2\,\phi/\phi^2\ \text{Cos}\,\phi. \tag{9}$$

Equation (9) is approximately of the form $\text{Sin}^5\,\phi/\phi^5$, and has almost exactly the form $\text{Sin}\,\phi/\phi$, if the scale in electrical degrees is adjusted for $r'''''=\sqrt{5}\cdot r$.

Thus, it may be induced as a working approximation, that aperture processes involving r indexes having values k times r, where r is the width index of the original picture element, are equivalent to a number of *identical* aperture processes, which number is approximately equal to k^2. Thus, scanning by a $2r$ aperture is equivalent to four successive scans by an r aperture.

If the number of identical aperture processes is K, then the rule of thumb is that the width index of the reproduced pulse is approximately

$$Kr = \sqrt{k_i^2 + k_1^2 + k_2^2 + \cdots k_p^2}\ r \tag{10}$$

where r is the width index of the original picture element, and k_i, $k_1 \cdots k_p$ are the ratios of aperture widths (in each aperture process) to picture-element width. (k_i is in itself equal to unity, by definition.)

Equation (10) is mathematically empirical, based on approximating a $(\text{Sin}\,\phi/\phi)^K$ curve with a $\text{Sin}\,\phi/\phi$ curve, plotted to a scale in electrical degrees, which scale has been contracted by the factor $1/K$.

The following theory assumes that optical systems are always composed of apertures which can be considered as a multitude of pinholes in parallel, each inclined to the other so that the transmitted images are coincident. The size of the pinholes (apertures) depends on the degree of optical excellence in grinding. The pinhole-aperture size would be difficult to specify, so it is more convenient to define resolution in terms of number of lines (usually, lines per millimeter, but in television practice, lines per frame).

In general, if an image of n_i lines is projected through successive optical systems having a resolving power of n_1, n_2, etc. lines, the over-all, approximate resolution of the reproduced image will be obtained by calculating the width index from (10) and converting this to a quality index, expressed as number of lines n_r in the reproduced image.

From (8) and (10) the quality formula may be obtained. Starting with an image of n_i line resolution, suppose that p different aperture processes, either scanning spot sizes or lens systems, are applied to obtain a reproduced image of n_r-lines resolution.

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[19] D. B. Smith, "Notes on the optimum number of lines for a television system," Panel 7, National Television Systems Committee.

[20] Figs. 2*b* to 2*e* and the curves of Fig. 6 are plotted in this manner. Plotting data are calculated in Appendix B.

[21] A filter of width r is defined as one which has a cutoff characteristic similar to that of an aperture of width r, as shown in curve 1 on Fig. 3, to the first elided frequency.

Then

$$\begin{aligned} k_i r &= 1/An_i^2 \\ k_1 r &= 1/An_1^2 \\ &\cdots\cdots \\ k_p r &= 1/An_p^2 \end{aligned} \tag{11}$$

and the reproduced resolution is

$$Kr = 1/An_r^2. \tag{12}$$

Equations (11) and (12) express the relation between the number of lines of resolution and the width of pulse for the optimum condition where the vertical and horizontal resolution are equal. This assumes that the number of lines is changed for the optimum condition each time the image passes through an aperture, which condition, of course, does not hold. Actually, the number of scanning lines is maintained constant at a value of n' lines per frame, so that the equations expressing the relationship between pulse width and the number of lines of resolution should be

$$\begin{aligned} &(\text{let } 1/rn'A = B) \\ k_i &= B/n_i \\ k_1 &= B/n_1 \\ &\cdots\cdots \\ k_p &= B/n_p \end{aligned} \tag{13}$$

and

$$K = B/n_r. \tag{14}$$

K is calculated by substituting from (13) in (10). Substitution for this value of K in (14) yields the approximate rule-of-thumb formula for over-all resolution

$$n_r = 1/\sqrt{1/n_i^2 + 1/n_1^2 + \cdots 1/n_p^2}. \tag{15}$$

An empirical formula for the peak values in Fig. 2 (or Appendix B) of the pulse shapes derived from the square pulse by subjecting it to x identical aperture processes is

$$\text{peak} = 1.18/\sqrt[4]{x(x+1)/2}. \tag{16}$$

Since the structure of the image itself is considered as one aperture process, the relative widening of a picture element after x-in-tandem aperture processes is the inverse ratio of (16) for values of $x=1$ and $x=x$ (inverse because width of pulse is inversely proportional to peak). The resolution of the reproduced image in terms of lines in the original image, for x identical aperture processes, thus may be expressed very accurately by

$$n_r = n_i/\sqrt[4]{x(x+1)/2}. \tag{17}$$

As an example, suppose $n_i = 800$ for a certain piece of film, $n_1 = 1500$ (30 lines per millimeter) for the projection lens on the camera photo-pickup tube, $n_2 = 600$ for the pickup-tube aperture, $n_3 = \infty$ for transmission filters (flat response in transmitter and receiver), $n_4 = 600$ for the projection-tube aperture (0.003-inch spot and 1.8-inch picture height), and $n_5 = 1000$ for the projection lens to the viewing screen. Substitution in (15) shows that the resolution on the viewing screen will be

$$(\text{without filters}) \qquad n_r = 342 \text{ lines}. \tag{18}$$

If the transmission filters are limited ($n_3 = 525$ lines instead of $n_3 = \infty$) the resolution will be degraded, and the line-structure will be

$$(\text{with filters}) \qquad n_r = 287 \text{ lines}. \tag{18a}$$

If 35-millimeter film is used, n_i becomes 1600, and the line structures become

$$(\text{without filters}) \qquad n_r = 368 \tag{19}$$

$$(\text{with filters}) \qquad n_r = 302. \tag{19a}$$

A comparison of (19) with (18) indicates, in the main, why 35-millimeter film has been found to give superior resolution to that given by 16-millimeter film in television practice. This superiority of 35-millimeter film over 16-millimeter film for television broadcasts can be expected to increase as the number of scanning lines is increased.

VI. Conclusions

It was shown in the last paragraph of Section IV that the widening effect of an aperture process, on an exact basis (integral method), is inversely proportional to the peak amplitude of the signal transmitted through the aperture. Thus, from the data given in Appendix B in regard to peak signals (values A_a, $A_{2a} \cdots A_{a4}$) the effects of successive aperture processes on the line structure of the reproduced image can be calculated in terms of per cent of original line structure. The last column in Table I was calculated according to the rule of thumb given in (15), whereby successive aperture processes will degrade the line structure of the original image such that the reproduced image will exhibit a certain percentage of the original structure.

Table I

Aperture Processes	Reproduced Lines n_r in Per Cent of Original	(From Rule of Thumb) n_r in Per Cent of Original
	Per Cent	Per Cent
Two identical	76.5	70.7
Three identical	64.2	57.7
Four identical	56.3	50.0
Five identical	42.8	44.4
Two processes, one twice the other		44.4
Two processes, one three times the other	28.1	31.6
Two processes, one four times the other	21.3	24.2

It can be seen that the percentage error introduced by the rule of thumb is of the order of ±10 per cent as compared to the more exact integral method, or the empirical formula (17) which is valid only for identical aperture processes.

In regard to the relative excellence of 35-millimeter film versus 16-millimeter film the resolution capability (line structure) of the film itself is undoubtedly a contributing factor of some importance, as shown by the calculations (18) and (19). There is another factor, beyond the scope of this paper, which affects the resolution in favor of 35-millimeter film. This factor is concerned with mechanical tolerances in regard to printing of film, registry of frames, projector jump, etc. These tolerances are more severe for 35-millimeter than for 16-millimeter film, which fact results in a still further improvement of resolution for 35-millimeter tele-motion pictures.

APPENDIX A

The Fourier analysis may be performed on any half cycle of the wave of Fig. 1, choosing the origin of coordinates at the center of a pulse, as shown.

Computation of the coefficients of the cosinusoidal terms in the Fourier series representing a finite pulse (sinusoidal terms are zero) is made as follows:

$$f(\phi) = bP \quad \text{from} \quad \phi = 0 \quad \text{to} \quad \phi = r\pi$$
$$f(\phi) = -rP \quad \text{from} \quad \phi = r\pi \quad \text{to} \quad \phi = \pi.$$

The coefficient of the Nth harmonic term is

$$b_n = \frac{2}{\pi}\int_0^{r\pi} bP \operatorname{Cos} n\phi \, d\phi + \frac{2}{\pi}\int_{r\pi}^{\pi} (-rP) \operatorname{Cos} n\phi \, d\phi. \quad (20)$$

n is an integer, and represents the order of a harmonic. By integration, and substitution of limits, (20) becomes

$$b_n = \frac{2P}{n\pi}(b \operatorname{Sin} nr\pi - r \operatorname{Sin} n\pi + r \operatorname{Sin} nr\pi) \quad (21)$$

so that

$$b_n = \frac{2P}{n\pi}(b+r) \operatorname{Sin} nr\pi = \frac{2P}{n\pi} \operatorname{Sin} nr\pi \ (\text{since } b+r=1). \quad (22)$$

Hence, the series is

$$A_i = 2rP \sum_{n=1}^{\infty} \frac{\operatorname{Sin} nr\pi}{nr\pi} \operatorname{Cos} n\omega t \quad (23)$$

where A_i defines the amplitude of the wave at any instant of time t.

It can be shown that the double-peak amplitude of (23) is

$$P = I\omega / br\pi \quad (24)$$

where the constant I represents the single-peak amplitude of the saw-tooth wave,[22] which is the integral of A_i.

APPENDIX B

PLOTTING DATA FOR FIG. 2a

Substitution of values of ϕ from tables in (5a) to (5f) will yield data which are sufficient to plot the curve of Fig. 2a for cutoff at first elided frequency where $N = 1/r$. It will be noticed that the periodicity of the overshoots corresponds to the cutoff frequency.

$$A_a = \frac{2P}{\pi} \operatorname{Si}(\pi) = 1.179P$$

$$A_b = \frac{P}{\pi}[\operatorname{Si}(\pi/2) + \operatorname{Si}(3\pi/2)] = 0.948P$$

$$A_c = \frac{P}{\pi} \operatorname{Si}(2\pi) = 0.451$$

$$A_d = \frac{P}{\pi}[\operatorname{Si}(5\pi/2) - \operatorname{Si}(\pi/2)] = 0.059P$$

$$A_e = \frac{P}{\pi}[\operatorname{Si}(3\pi) - \operatorname{Si}(\pi)] = -0.056P$$

$$A_f = \frac{P}{\pi}[\operatorname{Si}(4\pi) - \operatorname{Si}(2\pi)] = 0.023P.$$

Extrapolation from these formulas shows that for $\omega t = n\pi$, A_n can be generalized

$$A_n = \frac{P}{\pi}[\operatorname{Si}(n+1)\pi - \operatorname{Si}(n-1)\pi].$$

PLOTTING DATA FOR OTHER CURVES OF FIG. 2

(*I*) *For two identical aperture processes (Fig. 2b):*

$$A_{2a} = \frac{2P}{\pi} \operatorname{Si}(2\pi) = 0.903P$$

$$A_{2b} = \frac{P}{4\pi}[5 \operatorname{Si}(5\pi/2) + 3 \operatorname{Si}(3\pi/2) - 2 \operatorname{Si}(\pi/2)]$$
$$= 0.783P$$

$$A_{2c} = \frac{P}{2\pi}[3 \operatorname{Si}(3\pi) - \operatorname{Si}(\pi)] = 0.505P$$

A_{2d} (Not computed)

$$A_{2e} = \frac{2P}{\pi}[\operatorname{Si}(4\pi) - \operatorname{Si}(2\pi)] = 0.050P$$

$$A_{2f} = \frac{P}{2\pi}[5 \operatorname{Si}(5\pi) - 6 \operatorname{Si}(3\pi) + \operatorname{Si}(\pi)] = -0.006P.$$

(*II*) *For three identical aperture processes (Fig. 2c):*

$$A_{3a} = \frac{3P}{4\pi}[3 \operatorname{Si}(3\pi) - \operatorname{Si}(\pi)] = 0.758P$$

$$A_{3c} = \frac{P}{2\pi}[4 \operatorname{Si}(4\pi) - 2 \operatorname{Si}(2\pi)] = 0.498P$$

$$A_{3e} = \frac{P}{8\pi}[25 \operatorname{Si}(5\pi) - 27 \operatorname{Si}(3\pi) + 4 \operatorname{Si}(\pi)] = 0.121P$$

(*III*) *For four identical aperture processes (Fig. 2d):*

$$A_{4a} = \frac{2P}{3\pi}[4 \operatorname{Si}(4\pi) - 2 \operatorname{Si}(2\pi)] = 0.665P$$

(*IV*) *For two aperture processes; one of width 2πr; and one of width 4πr (corresponds to scanning with an aperture of twice the width of image picture element) (Fig. 2e):*

$$A_{a2} = \frac{P}{2\pi}[3 \operatorname{Si}(3\pi) - \operatorname{Si}(\pi)] = 0.505P$$

$$A_{c2} = \frac{P}{\pi} \operatorname{Si}(4\pi) = 0.475P$$

$$A_{e2} = \frac{P}{4\pi}[5 \operatorname{Si}(5\pi) - 3 \operatorname{Si}(3\pi)] = 0.251P$$

$$A_{f2} = \frac{P}{2\pi}[3 \operatorname{Si}(6\pi) - 2 \operatorname{Si}(4\pi) - \operatorname{Si}(2\pi)] = 0.023P$$

(*V*) *For two aperture processes; one of width 2πr; and one of width 6πr; (corresponds to scanning with aperture three times width of image picture element) (Fig. 2f):*

$$A_{a3} = \frac{2P}{3\pi}[2 \operatorname{Si}(4\pi) - \operatorname{Si}(2\pi)] = 0.332P.$$

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[22] T. E. Shea, "Transmission Networks and Wave Filters." D. van Nostrand Co., Inc., New York, N. Y., 1929.

(*VI*) *For two aperture processes; one of width $2\pi r$; and one of width $8\pi r$ (corresponds to scanning with aperture four times width of image picture element) (Fig. 2g):*

$$A_{a4} = \frac{P}{4\pi}\left[5 \text{ Si } (5\pi) - 3 \text{ Si } (3\pi)\right] = 0.251P.$$

Peak values of A_{xa} for any number of identical aperture processes x can be calculated approximately from the empirical formula (16)

$$A_{xa} = 1.18/\sqrt[4]{x(x+1)/2}.$$

APPENDIX C

In order that E_a of (6) should represent the *instantaneous* electrical response from the photocell, due to the motion of the aperture, it is necessary that the constant K', which includes the sensitivity of the photo surface and the impedance of the load, contain also the width of the aperture as a divisor. That is, the integrated response from A_i over the aperture, divided by the aperture width, represents the instantaneous signal. Thus

$$K' = \frac{K''}{\Delta\phi'}. \tag{25}$$

The width of the aperture divided by the total length of scanning path (2π radians) will be called the index of the aperture in relation to the scanning cycle

$$r' = \frac{\Delta\phi'}{2\pi} \qquad K' = \frac{K''}{2\pi r'}. \tag{26}$$

Let the phase of the pulse in the cycle be $\omega t = \phi$. Substitute K', ωt, and A_i in (6).

$$E_a = \frac{2K''rP}{2\pi r'} \int_{n(\phi-\Delta\phi'/2)}^{n(\phi+\Delta\phi'/2)} \sum_{n=1}^{\infty} \frac{\text{Sin } nr\pi}{nr\pi} \text{Cos } (n\phi) \frac{d(n\phi)}{n}. \tag{27}$$

(The variable of integration is ($n\phi$)). The integration of the cosine function is obtained easily, as

$$E_a = \frac{A'}{2\pi r'} \sum_{n=1}^{\infty} \frac{\text{Sin } nr\pi}{nr\pi} \left.\frac{\text{Sin } (n\phi)}{n}\right]_{n(\phi-\Delta\phi'/2)}^{n(\phi+\Delta\phi'/2)}$$

$$\text{(Where } A' = 2K''rP\text{)}. \tag{28}$$

Substituting limits

$$E_a = A' \sum_{n=1}^{\infty} \frac{\text{Sin } nr\pi}{nr\pi} \; \frac{2 \text{ Cos } n\phi \text{ Sin } \dfrac{n\Delta\phi'}{2}}{2\pi r' n}. \tag{29}$$

Substitute for $\Delta\phi'$ and ϕ: ($\Delta\phi' = 2\pi r'$, $\phi = \omega t$)

$$E_a = A' \sum_{n=1}^{\infty} \frac{\text{Sin } nr\pi}{nr\pi} \; \frac{\text{Sin } nr'\pi}{nr'\pi} \text{ Cos } n\omega t \tag{30}$$

which is (7) of the text.

APPENDIX D

To find the equation of the infinitesimal pulse, let r approach zero in (1).

$$A_i' = \lim_{r\to 0} A_i. \tag{31}$$

Since r is contained explicitly in P, it will be necessary to substitute for P in (31). (See (24), Appendix A).

$$A_i' = \lim_{r\to 0} \frac{2I\omega}{\pi} \sum_{n=1}^{\infty} \frac{\text{Sin } nr\pi}{nr\pi(1-r)} \text{ Cos } n\omega t. \tag{32}$$

Differentiate the numerator and denominator of (32)

$$A_i' = \left.\frac{2I\omega}{\pi} \sum_{n=1}^{\infty} \frac{n\pi \text{ Cos } nr\pi}{n\pi(-r + (1-r))} \text{ Cos } n\omega t \right]_{r=0}. \tag{33}$$

At the limit $r = 0$, the coefficient of Cos $n\omega t$ becomes unity, and the series becomes divergent.

$$A_i' = \frac{2I\omega}{\pi} \sum_{n=1}^{\infty} \text{Cos } n\omega t. \tag{34}$$

The method of Section III may be applied to (34) in order to obtain the shape of A_i' when only a finite number of harmonics represented by $N = 1/r$ is plotted. Multiply (34) by $\Delta\phi/r\pi$ (= unity), and replace the summation by an integral (this is allowable if r is small).

$$(A_i) \text{ finite} = \frac{2I\omega}{r\pi^2} \int_0^{\phi_1} \text{Cos } n\omega t \, d\phi \tag{35}$$

where $\phi_1 = Nr\pi$. (This is equivalent to summation from 1 to N instead of 1 to infinity. Limits are changed from n to ϕ. See Section III.)

At the center of the pulse, $\omega t = 0$, so

$$(A_i')_0 = \frac{2I\omega}{r\pi^2} \int_0^{\pi} d\phi = \frac{2I\omega}{r\pi} = U/r. \tag{36}$$

At $\omega t = r\pi/2$

$$(A_i')_{r/2} = \frac{2U}{\pi r} \int_0^{\pi} \text{Cos } (\phi/2) d(\phi/2) = 2U/\pi r = 0.636U/r. \tag{37}$$

At $\omega t = r\pi$

$$(A_i')_r = \frac{U}{\pi r} \int_0^{\pi} \text{Cos } \phi \, d\phi = 0. \tag{38}$$

At $\omega t = 3r\pi/2$

$$(A_i')_{3r/2} = \frac{2U}{3\pi r} \int_0^{\pi} \text{Cos } (3\phi/2) \; d(3\phi/2) =$$

$$-\,2U/3\pi r = -\,0.212U/r. \tag{39}$$

Thus, by expressing ωt in terms of $r\pi$, the amplitude of the infinitesimal pulse with finite cutoff at $1/r$ harmonics can be calculated at any phase point desired (see Fig. 4).

ACKNOWLEDGMENT

Grateful acknowledgment is made to Dr. Hans W. Salinger for his assistance in performing the integrations and in deriving some of the more difficult mathematical formulas of Appendix B.

Loop-Antenna Coupling-Transformer Design*

W. S. BACHMAN†, MEMBER, I.R.E.

Summary—**The low-impedance loop coupling-transformer circuit is analyzed, expressing the transformer parameters in terms of the circuit and transformer coupling coefficients. Equations are developed which yield optimum design values for the transformer. It is shown that an ideal transformer-coupled loop has 38.4 per cent of the gain realizable from a direct-connected loop of the same area, assuming the same Q in the transformer secondary as in the direct-connected loop.**

BUILT-IN loop antennas are widely used in broadcast receivers because of their convenience and simplicity from the consumer's point of view. In order to make them most effective, it is usual to shield them electrostatically against local disturbances. Considerable simplification is possible, if, instead of an electrostatically shielded high-impedance loop, a low-impedance loop with coupling transformer is used. This alternative permits greater tuning range because of the lower distributed capacitance of the system, and gives more flexibility in the choice of size and location of the loop.

A further simplification would be effected in all-wave receivers if the short-wave loop were also used as the low-impedance broadcast-band loop. This fixes the inductance of the loop at the necessary value for the short-wave band to be covered. Since the range of the tuning capacitor is also fixed, it is necessary to design a transformer to couple the loop to the tuning capacitor, and to specify optimum values for its parameters.

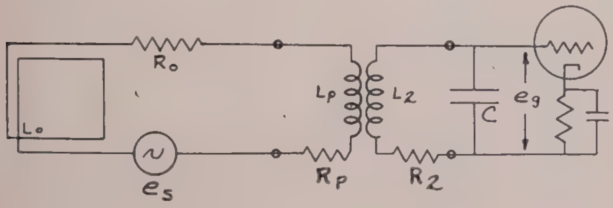

Fig. 1—Low-impedance loop coupling-transformer circuit.

In the circuit of Fig. 1, it is necessary to specify the values for L_p, the primary-winding inductance; L_2, the secondary inductance; and M, the mutual inductance.

Let

$$L_1 = L_0 + L_p$$

$$R_1 = R_0 + R_p$$

$$K = \frac{M}{\sqrt{L_1 L_2}} \tag{1}$$

$$K_1 = \frac{M}{\sqrt{L_p L_2}} \tag{2}$$

where K is the coupling coefficient between the entire primary circuit and the secondary, and K_1 is that between the primary winding only and the secondary. By expressing the unknowns in terms of K and K_1, and finding the optimum values for these coefficients, it is possible to specify the optimum values for the transformer parameters.

Analyzing the circuit of Fig. 1,

$$Z_1 = R_1 + jX_{L_1}$$

$$Z_2 = R_2 + j(X_{L_2} - X_c).$$

Then

$$I_1 = \frac{e_s}{Z_1 + \dfrac{X_m^2}{Z_2}}; \qquad e_s = I_1\left(Z_1 + \frac{X_m^2}{Z_2}\right)$$

$$I_2 = \frac{jX_m I_1}{Z_2}; \qquad e_g = -jX_c I_2 = \frac{X_c X_m I_1}{Z_2}.$$

The gain A is

$$A = \frac{e_g}{e_s} = \frac{\dfrac{X_c X_m}{Z_2}}{Z_1 + \dfrac{X_m^2}{Z_2}} = \frac{X_c X_m}{Z_1 Z_2 + X_m^2}. \tag{3}$$

Referring to the secondary circuit, in addition to X_{L_2} and X_c, there will be coupled in impedance from the primary as shown in Fig. 2.

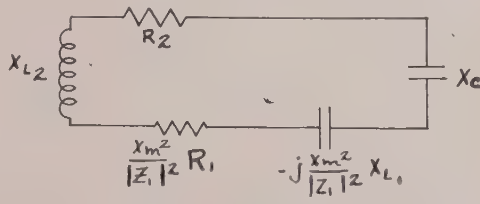
R_2
X_{L_2}
X_c
$\frac{X_m^2}{|Z_1|^2} R_1$
$-j\frac{X_m^2}{|Z_1|^2} X_{L_1}$

Fig. 2—Equivalent circuit, referred to the secondary, of the low-impedance loop coupling transformer.

At resonance the reactances balance out, so that

$$jX_{L_2} - j\frac{X_m^2}{|Z_1|^2} X_{L_1} - jX_c = 0. \tag{4}$$

Now if the primary Q is reasonably high,

$$X_{L_1} \gg R_1 \text{ and } |Z_1|^2 \text{is very nearly equal to } |X_{L_1}|^2.$$

Therefore (4) becomes

$$jX_{L_2} - j\frac{X_m^2}{X_{L_1}} - jX_c = 0$$

$$X_m^2 = X_{L_1}(X_{L_2} - X_c). \tag{5}$$

Expanding the expression for gain (3) and substituting (5) for X_m^2, we obtain

$$A = \frac{X_c X_m}{R_1 R_2 + j[R_1(X_{L_2} - X_c) + R_2 X_{L_1}]}.$$

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* Decimal classification: R382.1×R325.3. Original manuscript received by the Institute, June 20, 1945.

† General Electric Company, Bridgeport, Conn.

For reasonably high values of Q this reduces to

$$A = \frac{X_c X_m}{R_1(X_{L_2} - X_c) + R_2 X_{L_1}}. \tag{6}$$

If

$$Q_1 = \frac{X_{L_1}}{R_1} \quad \text{and} \quad Q_2 = \frac{X_{L_2}}{R_2}$$

(6) becomes

$$A = \frac{X_c X_m Q_1 Q_2}{Q_2 X_{L_1}(X_{L_2} - X_c) + Q_1 X_{L_1} X_{L_2}}. \tag{7}$$

From (1) we may write

$$X_m{}^2 = K^2 X_{L_1} X_{L_2}. \tag{8}$$

Equating (5) and (8)

$$X_{L_1}(X_{L_2} - X_c) = K^2 X_{L_1} X_{L_2}. \tag{9}$$

Substituting (9) in (7)

$$A = \frac{X_c X_m Q_1 Q_2}{(Q_1 + K^2 Q_2) X_{L_1} X_{L_2}}. \tag{10}$$

Substituting the value of X_m from (8) in (10)

$$A = \frac{X_c K}{\sqrt{X_{L_1} X_{L_2}}}\left(\frac{Q_1 Q_2}{Q_1 + K^2 Q_2}\right). \tag{11}$$

From (1) and (2) we obtain

$$X_{L_1} = \frac{X_{L_0}}{\left(1 - \dfrac{K^2}{K_1{}^2}\right)}. \tag{13}$$

Solving (9) for X_{L_2}

$$X_{L_2} = \frac{X_c}{(1 - K^2)}. \tag{14}$$

Substituting (13) and (14) in (11)

$$A = \frac{X_c K}{\sqrt{\left(\dfrac{X_{L_0}}{\left(1 - \dfrac{K^2}{K_1{}^2}\right)}\right)\left(\dfrac{X_c}{(1 - K^2)}\right)}}\left(\frac{Q_1 Q_2}{Q_1 + K^2 Q_2}\right)$$

$$A = \sqrt{\frac{X_c}{X_{L_0}}}\left(\frac{Q_1 Q_2}{Q_1 + K^2 Q_2}\right) \cdot \left[K^2(1 - K^2)\left(1 - \frac{K^2}{K_1{}^2}\right)\right]^{1/2}. \tag{15}$$

Equation (15) expresses the gain of the transformer in terms of two variables K and K_1 and the circuit Q's, and the unknown constants of the transformer are related to these values of K and K_1 by (13) and (14).

As an experimental verification of the theory, the following measurements were made on a developmental transformer:

$$\left.\begin{array}{rl} L_0 = & 7.0 \text{ microhenries} \\ L_p = & 3.6 \text{ microhenries} \\ L_2 = & 244.4 \text{ microhenries} \\ L_p + L_2 + 2M = & 284.9 \text{ microhenries} \end{array}\right\} \text{1000-cycle measurements}$$

$$\left.\begin{array}{rl} Q_0 = & 82 \\ Q_2 = & 153 \\ C = & 118 \text{ micromicrofarads} \end{array}\right\} \text{1000-kilocycle measurements.}$$

These inductance measurements yield values of 0.625 for K_1, and 0.364 for K.

$$X_c = \frac{1}{2\pi f C} = 1350 \text{ ohms at 1000 kilocycles}$$

$$X_{L_0} = 2\pi f L_0 = 44 \text{ ohms at 1000 kilocycles.}$$

Substituting in (15)

$$A = \sqrt{\frac{1350}{44}}\left(\frac{82 \times 153}{82 + 0.132 \times 153}\right)[0.132(0.868)(0.663)]^{1/2}$$

$$= 5.54 \times 123 \times 0.276 = 188.$$

This checks very closely with the measured gain of 190. The circuit gain was measured by inserting the source voltage of a Q meter in series with the primary circuit, and connecting the secondary across its capacitor terminals. The Q reading indicates the gain of the circuit, provided that the source impedance of the Q meter is small compared to the primary-circuit impedance. In this case, the source was 0.04 ohms, and the resistance in the primary circuit must be greater than 0.497 ohms, which is known to be the radio-frequency resistance of the loop itself at 1000 kilocycles. This indicates an error of something less than 7 per cent in the measured value.

Examination of (15) shows that the variable K_1 appears only once, and that it should be as large as possible for maximum gain. It should be noted that K must always be smaller than K_1, for, if they were equal, all of the primary-circuit inductance would be in the primary winding, leaving no inductance in the loop, and, hence, no signal pickup would be obtained.

Knowing that K_1 should be as large as possible, it is necessary to determine experimentally what value of K_1 may be practically attained. Values in the order of 0.9 have been readily attained in iron-cored transformers for this service.

As is seen from (15), the circuit Q has some effect on the best value of K. In order to simplify the Q parenthesis, it would be convenient to express Q_1 in terms of Q_2. Let

$$a = \frac{Q_1}{Q_2}; \qquad Q_1 = aQ_2.$$

Substituting this in (15)

$$A = \sqrt{\frac{X_c}{X_L}}\, Q_2\left(\frac{a}{a + K^2}\right)\left[K^2(1 - K^2)\left(1 - \frac{K^2}{K_1{}^2}\right)\right]^{1/2}. \tag{16}$$

Having this equation, it is necessary only to substitute known values to yield all of the necessary design information. The gain is directly proportional to Q_2, so the highest value practically realizable should be used. Having chosen Q_2, the ratio a may be estimated readily, since Q_1 is largely determined by the Q_0 of the loop itself. The value for K_1 has already been discussed. It is then possible to differentiate (16) with respect to K, and equate to zero, to find the optimum value for K.

Knowing the optimum K, it may be substituted in (13) and (14) to determine the required values of primary and secondary inductance.

Having arrived at the optimum design of the loop coupling transformer, it is of interest to compare the low-impedance transformer-coupled loop with a high-impedance loop connected directly to the tuning capacitor. The voltage induced into a loop is

$$e_s = \frac{2\pi NA}{\lambda} H \tag{17}$$

where N = the number of turns
A = the area of the loop
λ = the wavelength of the signal
H = the field strength.

The product $e_s\sqrt{X_c/X_{L_0}}$ may be shown to be independent of the number of turns, since

$$X_{L_0} = \omega L_0 = \omega N^2 P \tag{18}$$

where P is the permeance of the magnetic path. Substituting (18) and (17),

$$e_s \sqrt{\frac{X_c}{X_{L_0}}} = \frac{2\pi NAH}{\lambda} \sqrt{\frac{X_c}{\omega N^2 P}}.$$

In the case of the direct-connected high-impedance loop, $X_c = X_{L_0}$ hence the voltage at the grid is

$$e_g = \sqrt{\frac{X_c}{X_{L_0}}}\, e_s Q_2 = e_s Q_2.$$

In the low-impedance transformer-coupled loop

$$e_g = \sqrt{\frac{X_c}{X_{L_0}}}\, e_s Q_2 \left(\frac{a}{a+K^2}\right)\left[K^2(1-K^2)\left(1-\frac{K^2}{K_1{}^2}\right)\right]^{1/2}.$$

Since the term $\sqrt{(X_c/X_{L_0})}e_s$ has been shown to be independent of the number of turns, the expression

$$\left(\frac{a}{a+K^2}\right)\left[K^2(1-K^2)\left(1-\frac{K^2}{K_1{}^2}\right)\right]^{1/2} = \text{gain factor.}$$

This gain factor is the ratio of the gain of a transformer-coupled loop to a direct-connected loop working into the same tuning capacitance. This is always less than unity, and varies between 0.25 and 0.30 in practical cases.

In Fig. 3 are plotted several curves of gain factor versus K, for various combinations of primary Q and K_1. They show that the optimum value of K is not critical, and that it is not violently affected by the value of primary Q.

Of particular interest is the case of an infinite primary-circuit Q and a transformer coupling coefficient of unity. These values yield an optimum K of 0.576 ($K^2 = \frac{1}{3}$). Using this value in (13), it is found that L_0, the loop inductance, is $\frac{2}{3}$ of L_1, the primary-circuit inductance. The other $\frac{1}{3}$ is in the primary-winding inductance.

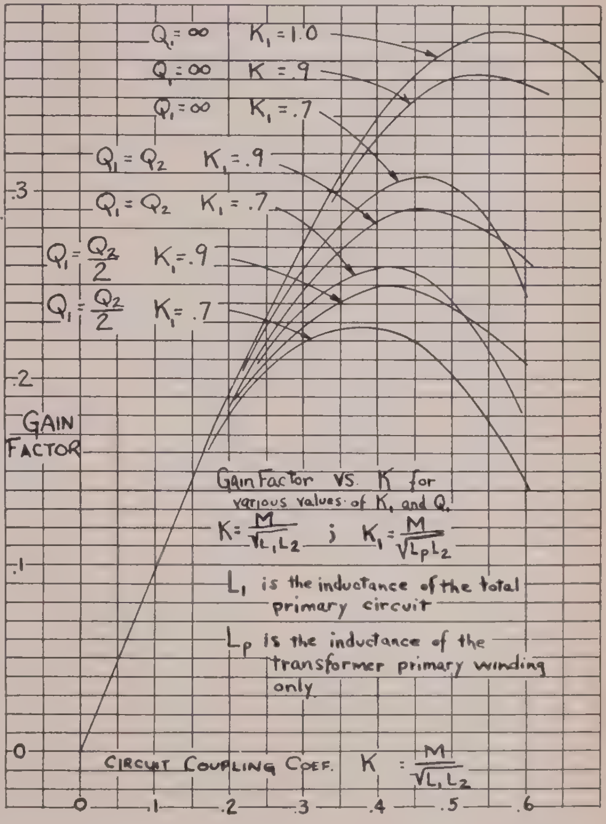

Fig. 3—Gain factor (ratio of the gain of a transformer-coupled loop working into the same tuning capacitance) versus circuit-coupling coefficient, for various values of primary circuit Q and transformer-coupling coefficient.

The gain factor, using this value of K, is 0.384, which is numerically equal to $\frac{2}{3} \times 0.576$. The physical significance of this is that only $\frac{2}{3}$ of the primary inductance provides the signal pickup, and since K is 0.576, only this fraction of the primary voltage would appear in the secondary of a 1:1 transformer. Thus, an ideal transformer-coupled loop antenna is capable of only 38.4 per cent of the gain of a direct-connected loop, assuming the secondary Q to be equal to the direct-connected loop Q, and equal loop areas in both cases. The distributed capacitance limits the usable size of direct-connected loops, so the low-impedance transformer-coupled loop may gain a considerable advantage in area. Also, it is quite often possible to obtain higher transformer-secondary Q than may be obtained in a high-impedance loop.

On the Theory of the Progressive Universal Winding*

A. W. SIMON†

Summary—**The rigorous theory previously developed by the author for the ordinary or stationary universal winding is extended to the progressive universal winding; in particular, an exact formula for calculating the gear ratio to be employed in winding a progressive universal coil is deduced, and, by neglecting certain terms therein, formulas representing various degrees of approximation of the same are derived and shown to be equivalent of those previously given by the author, by Joyner and Landon, and by Hershey. The theory of the spiral ridges and close-packed layers of progressive universal coils is developed, and formulas giving the slope of the spiral ridge and the dimensions of the close-packed layers are deduced. The principles underlying the selection of the optimum number of crossovers per turn are discussed, and an example of the application of the formulas to the design of a progressive universal coil is given.**

I. Theory of the Progressive Universal Winding

THE METHOD of winding a progressive universal coil is such that, as the dowel or tube on which the coil is wound rotates, the wire is guided back and forth by a shuttle, which displaces the wire in linear proportion to the angle of rotation of the dowel, while *simultaneously* the dowel is given a uniform linear motion in a direction parallel to its axis.

Table of Symbols

A, B, C a, b, c = constants appearing in the gear-ratio formulas; defined specifically in terms of the basic parameters by (9), (10), (11), (12), (13), and (14).

c = cam throw; i.e., maximum shuttle displacement produced by the cam.

δ = nominal diameter of the wire (including the insulation).

d = diameter of the dowel on which the coil is wound.

f = spacing factor; i.e., the ratio of the distance between centers of adjacent wires to the nominal wire diameter. ($f = s/\delta$)

h = circumferential advance per crossover; in particular, the distance which the point of crossover gains, or loses, over a line drawn, on the developed surface, at a distance from the starting line equal to the circumference divided by the approximate number of crossovers per turn. This can be easily seen in Figs. 1 and 2.

k = axial advance per crossover; the distance which the dowel moves in an axial direction during the time of one crossover.

n = approximate (integral) number of crossovers per turn of the dowel.

n' = exact number of crossovers per turn; equal to the exact number of half cam turns per turn of the dowel.

s_1 = spacing between centers of adjacent wires produced on the *forward* stroke.

s_2 = spacing between centers of adjacent wires produced on the *backward* stroke.

s_m = spacing of the points of maximum displacement of the wire in the direction of the spiral ridge.

s_r = spacing between successive turns of the (same) spiral ridge.

$t = \tan \phi$ (used in appendix only).

t = turns per inch (in general) when not otherwise indicated.

w_l = width of the close-packed layer.

w_o = width of the opening between close-packed layers.

x = linear displacement of a point of the wire on the developed surface in a direction *parallel* to the axis of the dowel.

y = linear displacement of a point of the wire on the developed surface in a direction *perpendicular* to the axis of the dowel.

θ_C = angle in radians through which the cam has turned.

θ_D = angle in radians through which the dowel has turned.

ϕ_1 = angle which the axis of the wire makes with the axis of the dowel on the *forward* stroke.

ϕ_2 = same for the *backward* stroke.

ψ = angle which the direction of the spiral ridge makes with the axis of the dowel.

ψ_0 = value of ψ for zero opening between close-packed layers.

Since the longitudinal motion or "progression" of the dowel alternately aids and opposes the cam motion, it is necessary in setting up the equations to consider the conditions obtaining both on the forward stroke (motions aiding) and on the backward stroke (motions opposing).

GLOSSARY

Cam throw: the shuttle displacement produced by one full stroke of the cam in one direction.

Crossover: a journey of the wire from a point of maximum displacement in one direction to that in the other, corresponding to one full stroke of the cam in one direction. One crossover corresponds also to *one-half* revolution of the cam.

Point of crossover: the point at which the wire reaches its maximum displacement in either direction (point corresponding to maximum shuttle displacement, point of reversal).

Progressive layering: the case where the point of crossover *advances* over a line drawn at a distance from the starting point equal to the length of the developed surface πd divided by the approximate number n of crossovers per turn ($h > 0$, Fig. 1).

Retrogressive layering: the case where the point of crossover falls *behind* the line defined under progressive layering. ($h < 0$, Fig. 2.)

Forward stroke: that corresponding to the case where the motion of the shuttle and the axial motion of the dowel are in the *same* direction.

Backward stroke: that corresponding to the case where the motion of the shuttle and the axial motion of the dowel are in *opposite* directions.

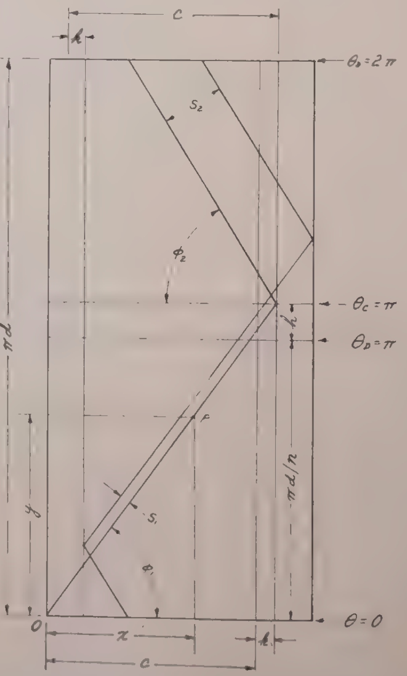

Fig. 1—Developed diagram of progressive universal winding with progressive layering. Two crossovers per turn.

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* Decimal classification: R 382. Original manuscript received by the Institute, April 25, 1945; revised manuscript received, August 3, 1945.

† Edwin I. Guthman and Company, Inc., Chicago, Illinois.

Accordingly (Figs. 1 and 2), if we focus attention on the linear displacement of a point P of the wire, on the developed surface, in a direction first parallel and then perpendicular to the axis of the dowel, we have

$$x/(c \pm k) = \theta_C/\pi \tag{1}$$

$$y/\pi d = \theta_D/2\pi \tag{2}$$

where the upper sign is taken for the forward stroke, and the lower for the backward stroke.

If we divide (2) by (1), and note that the quotient θ_D/θ_C is equal to the gear ratio r, there results

$$y/x = \pi dr/2(c \pm k). \tag{3}$$

If, next, we focus attention on the angles ϕ_1 and ϕ_2 which the wire makes with the axis of the dowel on the forward stroke and the backward stroke, respectively, we have

$$\tan \phi_1 = y/x = \pi dr/2(c+k) \tag{4a}$$

$$\tan \phi_2 = y/x = \pi dr/2(c-k). \tag{4b}$$

However, from Figs. 1 and 2, it is readily seen that these angles also satisfy the relations

$$(c+k) \tan \phi_1 = (\pi d/n) \pm h \tag{5a}$$

$$(c-k) \tan \phi_2 = (\pi d/n) \pm h \tag{5b}$$

where the upper sign is now taken for progressive layering,[1] and the lower for retrogressive layering, a convention *which will be adhered to in all that follows.*

As regards the spacing of adjacent wires (Fig. 3) we have

$$s_1 = qh \cos \phi_1 \mp qk \sin \phi_1 \tag{6a}$$

$$s_2 = qh \cos \phi_2 \pm qk \sin \phi_2. \tag{6b}$$

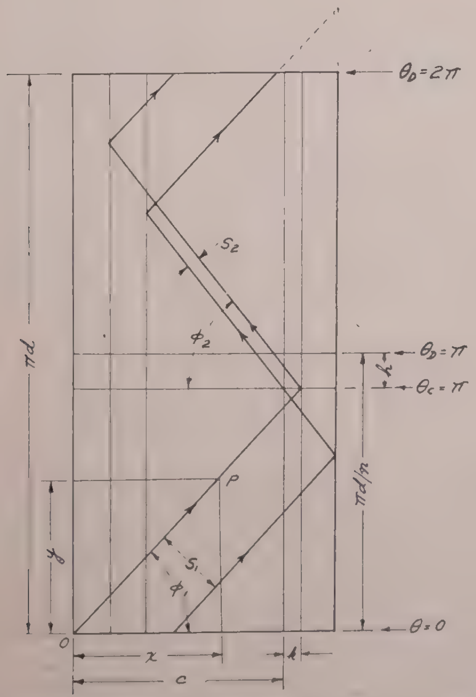

Fig. 2—Developed diagram of progressive universal winding with retrogressive layering. Two crossovers per turn.

In addition, the axial advance per crossover, the pitch of the progression, the *exact* number of crossovers per turn, and the gear ratio, are related according to the equations

$$k = p/n' = pr/2. \tag{7}$$

Since support is attained for progressive layering on the forward stroke and for retrogressive layering on the backward stroke, the elimination of ϕ_1 for the first and ϕ_2 for the second, from (4) and (6), after substituting therein for h and k their values as given by (5) and (7), respectively, yields the quadratic equation.

$$Ar^2 + Br + C = 0 \tag{8}$$

where the coefficients A, B, and C, are defined by

$$A = (1-a^2) \mp 2e + (1-b^2)e^2 \tag{9}$$

$$B = -(4/n)[1 \mp (1-b^2)e] \tag{10}$$

$$C = (4/n^2)(1-b^2) \tag{11}$$

and the constants a, b, and e, in turn by

$$a = s/qc \tag{12}$$

$$b = ns/q\pi d \tag{13}$$

$$e = p/nc. \tag{14}$$

Rigorous solution of (8) yields[2]

$$r = (2/n)\frac{[1 \pm (1-b^2)e \pm \sqrt{a^2+b^2-a^2b^2}]}{[(1-a^2) \mp 2e + (1-b^2)e^2]}. \tag{15}$$

Since a, b, and e are small quantities, the formula for r can be rewritten with a high degree of accuracy

$$r = (2/n)[1 \pm e \pm \sqrt{a^2+b^2} + a^2 - e^2] \tag{16}$$

and with still a fair degree of accuracy

$$r = (2/n)[1 \pm e \pm \sqrt{a^2+b^2}]. \tag{17}$$

Finally, since b ordinarily will be small compared with a, we can set *approximately*

$$r = (2/n)[1 \pm (e+a)]. \tag{18}$$

The last equation is, in fact, the exact equivalent of that given by Joyner and Landon,[3] and if we set therein $e=0$, that of Hershey.[4]

Finally, if we set $e=0$ in (16), we obtain the accurate formula previously given by the present author for ordinary or "stationary" universal coils.[5,6]

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[1] The use of the word "progressive" to describe the direction in which the layering proceeds should not be confused with its use to describe the type of winding.

[2] The details of the mathematical solution are outlined in the Appendix.

[3] A. A. Joyner and V. D. Landon, "Theory and design of progressive universal coils," *Communications*, vol. 18, pp. 5–8; September, 1938.

[4] L. M. Hershey, "The design of the universal winding," Proc. I.R.E., vol. 29, pp. 442–446; August, 1941.

[5] A. W. Simon, "Winding the universal coil," *Electronics*, vol. 9, pp. 22–24; October, 1936; Errata, p. 52; November, 1936.

[6] A. W. Simon, "On the winding of the universal coil," Proc. I.R.E., vol. 33, pp. 35–37; January, 1945.

II. GEOMETRY OF THE SPIRAL RIDGES AND LAYERS

As pointed out by Joyner and Landon, the locus of the points of maximum displacement of the wire forms a ridge traveling in the shape of a spiral around the coil, the direction of the spiral being that of a left-hand screw for progressive layering, and of a right-hand screw for retrogressive layering (compare Figs. 1, 2, and 3).

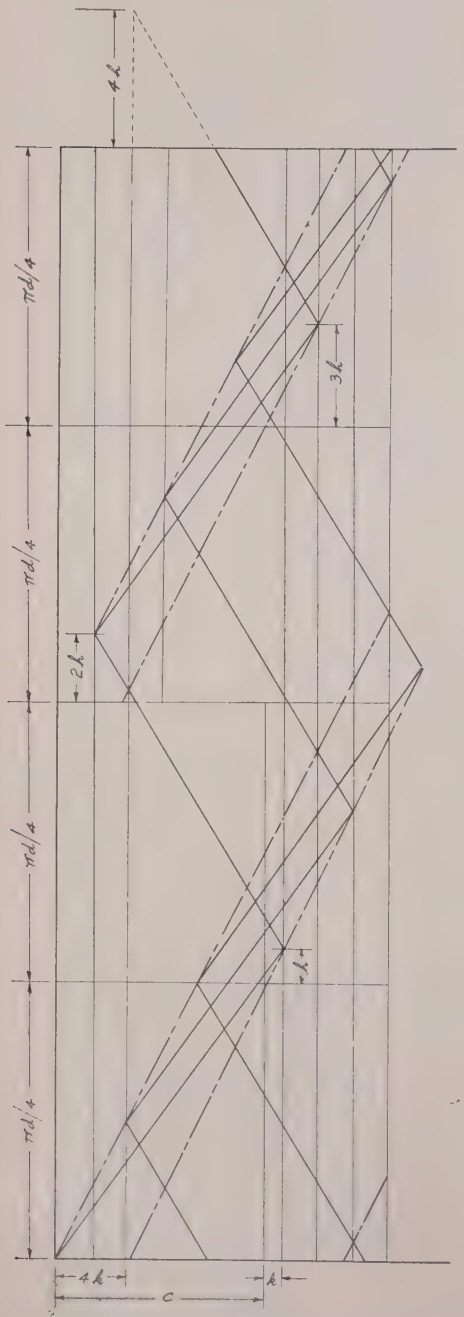

Fig. 3—Developed diagram of progressive universal winding with progressive layering. Four crossovers per turn. Two distinct close-packed layers.

If we denote by ψ the angle which the direction of this ridge makes with the axis of the coil, we have from Fig. 4 directly

$$\tan\psi = \pm h/k \tag{19}$$

or, in view of (7)

$$\tan\psi = \pm 2h/pr. \tag{20}$$

If next we eliminate h, from (4) and (6) we have

$$\pm h = \pi d(nr - 2)/2n \tag{21}$$

which value substituted in (20) yields

$$\tan\psi = \pm \pi d(1 - 2/nr)/p. \tag{22}$$

The last equation enables the calculation of the slope of the spiral ridge, provided the gear ratio r and the progression p are given.

Also, it is readily seen from Fig. 4 that the spacing s_m of the points of maximum displacement of the wire in the direction of the ridge is given by the simple formula

$$s_m = q\sqrt{h^2 + k^2} \tag{23}$$

while from Fig. 5 the spacing s_r between successive turns of the (same) ridge is given by

$$s_r = \pi d \cos\psi \tag{24}$$

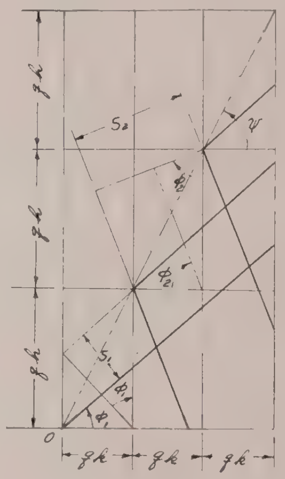

Fig. 4—Geometry of the wire spacing. Progressive layering.

and the width w_e of the close-packed layer is given by

$$w_e = c \sin\psi - (\pi d/n)\cos\psi; \tag{25}$$

consequently, the width w_0 of the opening between close-packed layers is

$$w_o = (1 + 1/n)\pi d \cos\psi - c \sin\psi. \tag{26}$$

Theoretically, then, the amount of this opening becomes zero when the ratio h/k is so chosen that

$$\tan\psi_0 = h/k_0 = (1 + 1/n)\pi d/c. \tag{27}$$

III. SELECTION OF THE NUMBER OF CROSSOVERS PER TURN

It should be pointed out that all the symbols used in the present paper have exactly the same significance as those used in the previous papers by the author, the only new concepts introduced being those relating to the progression; namely, the quantities k, p, t, and e. Hence we can take over, with only slight modification, the principles developed for the ordinary universal winding and apply the same to the progressive universal winding.

In particular, this holds for the selection of the number of crossovers per turn, which, as pointed out in a previous paper by the author, is selected on the basis that the winding angle ϕ have its minimum permissible value at the surface of the dowel, for the purpose of building up the coil as high as possible. Since the latter is not a necessary condition in the case of a progressive universal coil, the number of crossovers per turn can be selected so that the angle ϕ has a value which will assure absolutely no slippage. According to Joyner and Landon, this condition is attained when n satisfies the formula

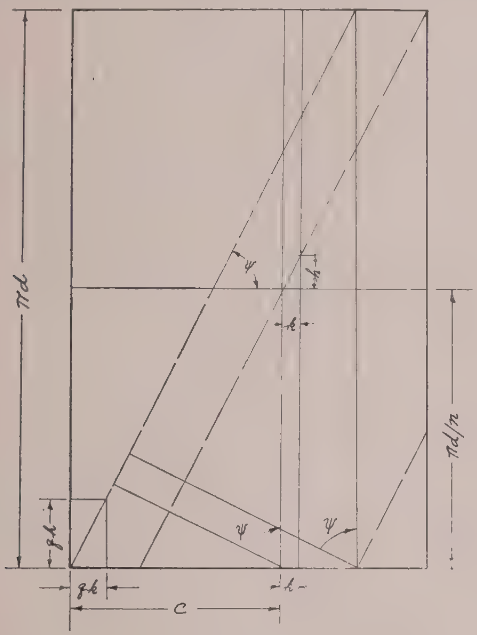

Fig. 5—Developed diagram of the spiral ridges and layers.

$$n = 2d/5c. \tag{28a}$$

The maximum number of crossovers per turn that can safely be used is given by

$$n = 2d/3c. \tag{28b}$$

IV. AN EXAMPLE OF THE APPLICATION OF THE FORMULAS

Let it be desired to wind a progressive universal coil with No. 32 S.S.E. wire ($\delta=0.0106$), on a $\frac{7}{8}$-inch form ($d=0.875$), using a $\frac{1}{8}$-inch cam ($c=0.125$), 120.5 turns per inch ($t=120.5$, $p=0.0083$), a spacing factor of 1.25, and retrogressive layering.

The maximum number of crossovers which can safely be used is

$$n = \frac{2 \times (\frac{7}{8})}{3 \times (\frac{1}{8})} = \frac{14}{8} \times \frac{8}{3} = 4\tfrac{2}{3}$$

while the condition of no slippage would give

$$n = \frac{2 \times (\frac{7}{8})}{5 \times (\frac{1}{8})} = \frac{14}{8} \times \frac{8}{5} = 2\tfrac{4}{5}$$

hence four crossovers per turn ($n=4$), corresponding to four crossovers per winding cycle ($q=4$) would constitute good design.

The parameters a^2, b^2, e, and e^2 are

$$\begin{aligned} a^2 &= (1.25 \times 0.0106/4 \times 0.125)^2 = 7.02 \times 10^{-4} \\ b^2 &= (4 \times 1.25 \times 0.0106/4 \times \pi \times 0.875)^2 \\ &= 2.32 \times 10^{-5} \\ e &= (0.0083/4 \times 0.125) = 1.66 \times 10^{-2} \\ e^2 &= (1.66 \times 10^{-2})^2 = 2.76 \times 10^{-5} \\ \sqrt{a^2 + b^2} &= 2.69 \times 10^{-2}. \end{aligned}$$

The accurate formula (16) then gives for r

$$\begin{aligned} r &= (2/4)(1.000 - 1.66 \times 10^{-2} - 2.69 \times 10^{-2} \\ &\qquad + 7.02 \times 10^{-4} - 2.76 \times 10^{-4}) \\ r &= (1/2)(1.000 - 4.3 \times 10^{-2}). \end{aligned}$$

The nearest integral ratio satisfying this equation is

$$r = (1/2)(1 - 1/23) = 22/46 = 44/92.$$

Hence a drive gear with 92 teeth and a cam gear with 44 teeth should be used.

APPENDIX

METHOD OF SOLUTION OF THE BASIC EQUATIONS

If we denote $\tan \phi$ by t, we have from (4) and (7)

$$t = \pi dr/(2c \pm pr) \tag{29}$$

while from trigonometry we have

$$\sin \phi = t/\sqrt{1 + t^2} \tag{30}$$

$$\cos \phi = 1/\sqrt{1 + t^2}. \tag{31}$$

Solving next (4) and (5) for h we obtain

$$\pm h = \pi d(nr - 2)/2n \tag{32}$$

while for k we have from (7)

$$k = pr/2. \tag{33}$$

If now in the basic (6) we substitute for $\sin \phi$, $\cos \phi$, h, and k, their values as given by (30) to (33) inclusive, we obtain

$$(2ns/q\pi d)\sqrt{1 + t^2} = (nr - 2) \mp nprt/\pi d. \tag{34}$$

If, finally, we substitute in (34) for t its value as given by (29), square, and rearrange terms, we obtain the basic quadratic (8), which is solved by the standard method.

Concerning Hallén's Integral Equation for Cylindrical Antennas*

S. A. SCHELKUNOFF†, FELLOW, I.R.E.

Summary—**The main purpose of this paper is to explain the substantial quantitative discrepancy between Hallén's formula for the impedance of cylindrical antennas, and ours. Hallén's first approximation involves a tacit assumption that the antenna is short compared with the wavelength. Since the subsequent approximations depend on the first, they are degraded by this initial assumption.**

The approximations involved in his integral equation itself are justified; and, if properly handled, the equation yields results in much better agreement with ours. The last section of the paper is devoted to infinitely long antennas. Such antennas can be treated by at least three very different methods and a comparison is instructive. In practice, the solution for this case is an approximation to a long antenna designed to carry progressive waves.

INTRODUCTION

IN A previous paper,[1] attention has been called to substantial discrepancies between the values of the impedance of cylindrical antennas as calculated from our formula[2] on the one hand, and from Hallén's formula on the other.[3] Since our results enjoy experimental support, it has been suggested,[4] tentatively, that the approximations involved in Hallén's integral equation from which his formula was obtained might account for the inaccuracy of his results. Our analysis shows, however, that these approximations are permissible under the contemplated conditions, and cannot account for the large discrepancies under consideration. On the other hand, Gray's solution[5] of Hallén's integral equation leads to an impedance formula which agrees far better with our formula than with Hallén's, see Table I.

All solutions under consideration, aside from mathematical details, are in the form of sequences of successive approximations, each term depending on the preceding. The goodness of the first approximation affects the goodness of the n-th. In all three solutions, only the first two approximations have actually been expressed in terms of known and tabulated functions. We find that the leading term of Hallén's series is obtained on a *tacit* assumption that the antenna is short compared with the wavelength. Hallén assumes explicitly that the parameter $\Omega = 2 \log (2l/a)$ is large (l is the length of one arm of the antenna and a is the radius); but in obtaining the magnitude of the leading term, he also assumes that l/λ is small and thus places a greater burden on the following correction terms. It is for this reason that Hallén's second approximation is inferior to ours, which does not assume that l/λ is small. By numerical integration, Bouwkamp obtained[6] the third term of Hallén's expansion and then computed a few values of the antiresonant impedance ($l \cong \lambda/2$) of center-fed cylindrical antennas in free space. Table I presents a comparison between various experimental and theoretical values. It is to be noted that the addition of the third term of Hallén's series alters the results by a large percentage in the direction of our values. There is no assurance, however, that the higher-order approximations will ever yield as good results for antennas of practicable dimensions as the first two terms of either Gray's formula or ours. Thus, Bouwkamp observes that the coefficients of the third term are already much larger than those of the second, and there is every indication that the trend will persist and that the series is not convergent but asymptotic. The added terms would not then improve the results unless the parameter Ω is much larger than it can possibly be for practical antennas. An antenna in Copenhagen is perhaps the "thinnest" tower on record; for this antenna, $\Omega = 20$. The diameter of the thinnest antenna measured by Feldman, at the Bell Telephone Laboratories, was 10 mils, and the length of one arm was about 30 feet; for this antenna, $\Omega = 23.8$. Doubling the length

TABLE I

TABLE OF MAXIMUM RESISTANCE OF A FULL-WAVE ANTENNA IN FREE SPACE ($l \cong \lambda/2$)

$\Omega = 2 \log (2l/a) = K_a/60 + 2$

K_a	$l/2a$	Ω	Theoretical				Experimental
			1	2	3	4	5
480	37.1	10	1620	1150	882*	680	860
780	452.0	15	3770	2900	2340	2220	2450
1080	5507.0	20	6820	5500	4840	4650	4840

(1) Ronold King and F. G. Blake, from two terms of Hallén's series;
(2) C. J. Bouwkamp, from three terms of Hallén's series;
(3) S. A. Schelkunoff, from two terms of his series;
(4) M. C. Gray, from two terms of her series;
(5) From an experimental* curve.

* In obtaining this value, the capacitance of the flat ends of the cylinder was estimated as explained in S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand and Co., Inc., New York, N. Y., 1943, p. 465. The *curves* in footnote reference 2 do not include this capacitance, which is negligible when $\Omega = 15$ or greater. The latter is concentrated near the rim and the removal of the flat ends would not appreciably alter the "end capacitance," particularly because the inner surface of the cylinder will become exposed. The same conclusion was reached experimentally by George H. Brown and O. M. Woodward, Jr., "Experimentally determined impedance characteristics of cylindrical antennas," PROC. I.R.E., vol. 33, pp. 257–262; April, 1945.

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* Decimal classification: R120. Original manuscript received by the Institute, June 18, 1945.

† Bell Telephone Laboratories, New York, N. Y.

[1] S. A. Schelkunoff, "Antenna theory and experiment," *Jour. Appl. Phys.*, vol. 15, pp. 54–60; January, 1944.

[2] S. A. Schelkunoff, "Theory of antennas of arbitrary size and shape," PROC. I.R.E., vol. 29, pp. 493–521; September, 1941.

[3] E. Hallén, "Theoretical investigations into the transmitting and receiving qualities of antennas," *Nova Acta* (Uppsala) [4] vol. 11, pp. 1–44; November, 1938.

[4] L. Brillouin, "The antenna problem," *Quart. Appl. Math.*, vol. 1, pp. 201–214; October, 1943.

[5] Marion C. Gray, "A modification of Hallén's solution of the antenna problem," *Jour. Appl. Phys.*, vol. 15, pp. 61–65; January, 1944.

[6] C. J. Bouwkamp, "Hallén's theory for a straight perfectly conducting wire, used as a transmitting or receiving aerial," *Physica*, vol. 9, pp. 609–631; July, 1942.

or halving the diameter would increase Ω by only 1.4; even $\Omega = 30$, a value which does not appear to be excessively large, can be of academic interest only since a 10-mil wire would have to be 1362 feet long.

In dealing with asymptotic series for relatively small values of the independent variable, it is quite possible to encounter a situation which is best illustrated by an example. The following two series belong to the same function

$$f(x) = \frac{10}{x} - \left(105 + \frac{5}{8}\right)\frac{1}{x^2} + \left(1118 + \frac{23}{64}\right)\frac{1}{x^3} - \cdots \quad \text{(A)}$$

$$= \frac{10}{x+10} - \frac{45}{8(x+10)^2} + \frac{375}{64(x+10)^3} - \cdots . \quad \text{(B)}$$

Both series are asymptotic in the sense that as x becomes larger and larger, any *fixed* number of terms will approximate $f(x)$ better and better; on the other hand, if x is not sufficiently large, the two series behave quite differently. Thus, if $x = 2$, the first three terms of series (B) yield $f(2) = 0.79766$, while the exact value of $f(2)$ is known to be 0.79728; but series (A) breaks down altogether. The exact value happens to be known because series (B) was obtained from

$$f(x) = 80 - 80e^{x+10}\sqrt{2(x+10)/\pi}\, K_0(x+10)$$

with the aid of the well-known asymptotic expansion for the modified Bessel function. Series (A) is nothing but (B), when the latter is expanded in powers of $1/x$ in place of $1/(x+10)$.

For the purposes of illustration, we have deliberately exaggerated the phenomenon which is actually encountered in the antenna theory. Nevertheless, the choice of the parameter in reciprocal powers of which the solution is expanded is much more important than is realized by some students of the theory.

Derivation of the Integral Equation for Antennas

Hallén's approximate integral equation for a *cylindrical antenna* is

$$\int_{-l}^{l} \frac{I(\xi)e^{-i\beta r}}{4\pi r}\, d\xi = A \cos \beta z + B \sin \beta |z|, \qquad (1)$$
$$r = \sqrt{(\xi - z)^2 + a^2}, \quad \beta = \omega\sqrt{\mu\epsilon} = 2\pi/\lambda,$$

where $I(z)$ is the current, l is the length of one arm of the antenna, a is the radius, and z is measured along the antenna from the generator. It should be emphasized that the above equation is for a cylindrical antenna, and that the equation changes with any change in the shape of the antenna. A recent suggestion[7] by King and Harrison that Hallén's solution of (1) actually applies to a pair of ellipsoids placed end to end, does not seem to take this fact into consideration.

The constants A and B are calculated from the impressed voltage and the condition imposed on the current at the ends of the antenna; but first, let us compare this equation with the exact equation for hollow, perfectly conducting cylindrical antennas. In order to obtain the exact equation, we need only observe that *if the currents on one or more parallel conductors are longitudinal, then on the surface of any conductor the retarded scalar and vector potentials are sinusoidal functions of the distance along the conductor.*[2,6] On a single cylinder, the longitudinal current flow is insured if the cylinder has no flat ends and if the impressed voltage is independent of the angular co-ordinate. For thin cylinders, the effects of the capacitance of these ends are small (of order a/l), and we are justified in neglecting them. Those who are interested in the exact equation for cylinders *with* the flat ends are referred to a paper by L. Brillouin.[4]

If V is the scalar retarded potential and Π the only nonvanishing component of the vector potential, then on the surface of the antenna and in the region free from the applied voltage, we have

$$\frac{dV}{dz} = -i\omega\mu\Pi, \qquad \frac{d\Pi}{dz} = -i\omega\epsilon V, \qquad (2)$$

where z is the axial co-ordinate. In the region of the impressed voltage the first equation becomes

$$\frac{dV}{dz} = -i\omega\mu\Pi + E^i(z), \qquad (3)$$

where $E^i(z)$ is the impressed voltage per unit length. Eliminating V, we obtain

$$\frac{d^2\Pi}{dz^2} = -\beta^2\Pi - i\omega\epsilon E^i(z). \qquad (4)$$

Let us now apply these equations to a pair of conducting cylinders, separated by a gap of length s, Fig. 1. Under the influence of the impressed voltage, the charge surges back and forth between the cylinders. This is not precisely the way antennas are operated, but the results can be adapted to the actual method of feeding shown in Fig. 2. Outside the gap, Π is sinusoidal, and we have

$$\Pi = \int_0^{2\pi}\int_{-l}^{l} \frac{I(\xi)e^{-i\beta r}}{8\pi^2 r}\, d\xi d\phi$$
$$= A \cos \beta z + B \sin \beta |z|, \quad |z| > s/2, \qquad (5)$$

where $I(z)$ is the current; also

$$r = \sqrt{(\xi - z)^2 + \rho^2}, \qquad \rho = 2a \sin(\phi/2), \qquad (6)$$

where a is the radius of the cylinder. It has been assumed that the impressed voltage is symmetric about the center; otherwise, we should have two equations of form (5), one for each arm of the antenna, with different values of A and B.

For a *uniform* impressed intensity across the gap on the continuation of the antenna surface we have[8]

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[7] Ronald King and C. W. Harrison, Jr., "The impedance of short, long, and capacitively loaded antennas with a critical discussion of the antenna problem," *Jour. Appl. Phys.*, vol. 15, pp. 170–185; February, 1944.

[8] This assumes that the internal impedance of the "generator" is zero. There is no loss in generality in this since the generator impedance is simply in series with the antenna impedance.

$$\Pi = C \cos \beta z + V^i/i\omega\mu s, \tag{7}$$

where $V^i = E^i s$. The retarded scalar potential is obtained by differentiation

$$\begin{aligned} V = -\frac{1}{i\omega\epsilon}\frac{d\Pi}{dz} &= -i\eta A \sin \beta z + i\eta B \cos \beta z,\ z > s/2; \\ &= -i\eta C \sin \beta z,\ -s/2 < z < s/2; \\ &= -i\eta A \sin \beta z - i\eta B \cos \beta z,\ z < -s/2; \end{aligned} \tag{8}$$

where

$$\eta = \sqrt{\mu/\epsilon}.$$

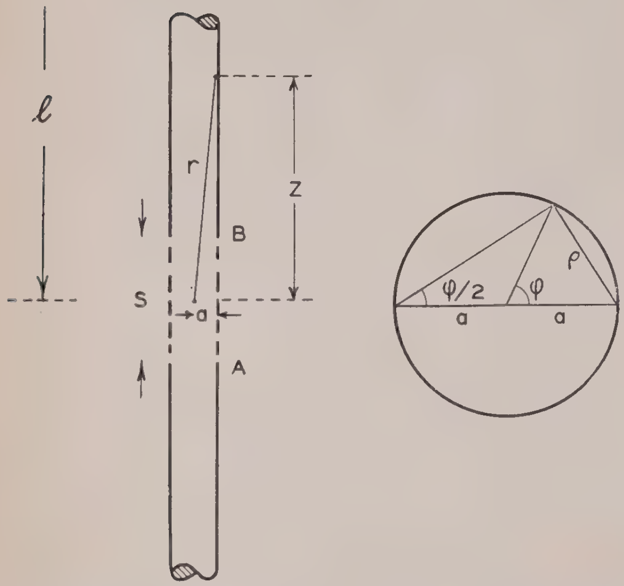

Fig. 1—A portion of a cylindrical antenna in the vicinity of the gap across which the electric charge is forced to flow back and forth between the two arms of the antenna.

The continuity of V and Π at $z = s/2$ gives two equations for determination of the constants of integration A, B, C. The continuity at $z = -s/2$ is automatically provided by the symmetry of the equations. At the ends of the antenna the current vanishes

$$I(\pm l) = 0. \tag{9}$$

Hence we have enough equations for obtaining the constants of integration. The ratio

$$Z_i = V^i/I(s/2) \tag{10}$$

is the input impedance of the antenna.

The input impedance depends, to a slight extent, on the distribution of the impressed voltage. This ambiguity is inevitable for a finite gap. For an infinitesimal gap the ambiguity is removed but the impedance necessarily vanishes, unless the "terminals" of the antenna are tapered as a pair of cones. In the immediate vicinity of an infinitesimal gap between two cylinders, the dominant term in the expression[2] for the current is $4i\omega\epsilon a V^i \log (a/z) + \text{const.} = (ia/15\lambda) V^i \log (a/z) + \text{const.}$ The first term is large only so long as z is very much smaller than a; still at $z=0$, the current is infinite and in impedance calculations we must assume a finite gap unless a approaches zero together with z sufficiently fast.

An alternative to assuming a distribution of impressed voltage is *to assume a constant current I between the terminals A and B and to calculate the voltage necessary to produce it.* In this case, we have to solve (5), (9) and

$$I(\pm s/2) = I. \tag{11}$$

Constants A and B will thus be obtained from (9) and (11). The necessary impressed voltage is then obtained by integration of (3)

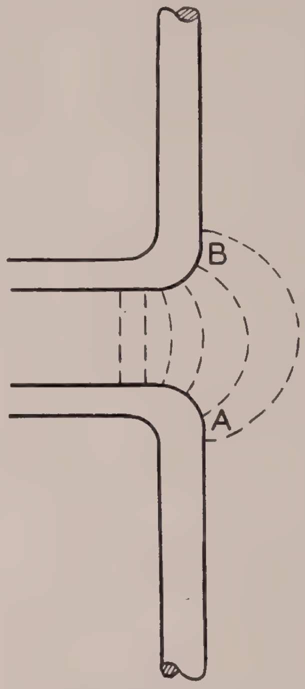

Fig. 2—A portion of a cylindrical antenna receiving its power from a two-wire transmission line.

$$\begin{aligned} V^i = \int_{-s/2}^{s/2} E^i(z)dz &= V(s/2) - V(-s/2) + i\omega\mu \int_{-s/2}^{s/2} \Pi(z)dz \\ &= 2V(s/2) + i\omega\mu \int_{-s/2}^{s/2} \Pi(z)dz \end{aligned} \tag{12}$$

with a subsequent substitution from (8). The main part of the last term is contributed by the vector potential of the current in the filament AB; hence, the ratio of this term to the current I represents largely the impedance of the filament. In order to obtain the impedance of the antenna as seen from the terminals A, B, this ratio should be subtracted from V^i/I; thus

$$Z_i = 2V(s/2)/I. \tag{13}$$

This is a better expression for the antenna impedance than (10), but usually there will be little numerical difference between the two.

CONCERNING THE SOLUTION OF THE INTEGRAL EQUATION FOR ANTENNAS

Let us now apply to the left side of (5) a transformation analogous to that employed by Hallén in connection with (1); thus we write

$$\int_0^{2\pi}\int_{-l}^{l}\frac{I(\xi)e^{-i\beta r}}{2\pi r}d\xi d\phi = I(z)\int_0^{2\pi}\int_{-l}^{l}\frac{d\xi d\phi}{2\pi r} + \int_0^{2\pi}\int_{-l}^{l}\frac{I(\xi)e^{-i\beta r}-I(z)}{2\pi r}d\xi d\phi. \quad (14)$$

The purpose of this resolution is to set the stage for successive approximations. If a is small and ξ is in the vicinity of $\xi = z$, r is also small; hence, as a approaches zero the first term on the right becomes infinite while the second remains bounded. Neglecting the second term, we obtain the first approximation for $I(z)$ by substitution in (5). Successive corrections are obtained by substituting each preceding approximation to $I(z)$ in the second term of (14). The integral

$$\Omega(z) = \int_0^{2\pi}\int_{-l}^{l}\frac{d\xi d\phi}{2\pi r} \quad (15)$$

itself is a fairly complicated function of z, and since we have to divide the right-hand side of (5) by this function, subsequent integrations of the second term in (14) would be complicated. To simplify the calculations, Hallén writes

$$\Omega(z) = \Omega(0) + [\Omega(z) - \Omega(0)]. \quad (16)$$

Instead of $\Omega(0)$ one could use the mean value of $\Omega(z)$, but the difference between these values turns out to be small. With the aid of (14) and (16), (5) becomes

$$I(z) = \frac{4\pi[A\cos\beta z + B\sin\beta|z|]}{\Omega(0)} + \left[1 - \frac{\Omega(z)}{\Omega(0)}\right]I(z) + \frac{1}{\Omega(0)}\int_0^{2\pi}\int_{-l}^{l}\frac{I(\xi)e^{-i\beta r}-I(z)}{2\pi r}d\xi d\phi. \quad (17)$$

Letting

$$I_0(z) = \frac{4\pi[A\cos\beta z + B\sin\beta|z|]}{\Omega(0)}, \quad (18)$$

substituting in the two last terms of (17), and continuing the process, we obtain

$$I(z) = I_0(z) + I_1(z) + I_2(z) + \cdots. \quad (19)$$

Evidently, this is a series in descending powers of $\Omega(0)$. In Hallén's analysis

$$\Omega(0) = 2\log(2l/a), \quad (20)$$

because he uses (1) rather than (5) and thus assumes in effect that $\rho = a$ in (15). It is this approximation that was criticized by Brillouin. It can be shown, however, that the error is of the order $(a/l)^2$. Integrating (15) with respect to ξ, we have

$$\Omega(z) = \frac{1}{2\pi}\int_0^{2\pi}\log\frac{[(l+z)+\sqrt{(l+z)^2+\rho^2}][(l-z)+\sqrt{(l-z)^2+\rho^2}]}{\rho^2}d\phi. \quad (21)$$

At points far enough from the ends we have

$$\begin{aligned}\Omega(z) &= \frac{1}{2\pi}\int_0^{2\pi}\log\frac{l^2-z^2}{a^2\sin^2(\phi/2)}d\phi \\ &\quad + \frac{1}{2\pi}\int_0^{2\pi}\left[\frac{a^2}{(l+z)^2}+\frac{a^2}{(l-z)^2}\right]\sin^2(\phi/2)d\phi + \cdots \\ &= \log\frac{4(l^2-z^2)}{a^2} + \frac{1}{2}\left[\frac{a^2}{(l+z)^2}+\frac{a^2}{(l-z)^2}\right] + \cdots \quad (22)\end{aligned}$$

where the omitted terms are of the order of a^4. At $z = 0$ this equation differs from (20) by a^2/l^2. On the surface, it may appear that Hallén's approximation is not justified near the ends of the wire; but it is. At $z = l$, we have

$$\begin{aligned}\Omega(l) &= \frac{1}{2\pi}\int_0^{2\pi}\log\frac{2l+\sqrt{4l^2+\rho^2}}{\rho}d\phi \\ &= \frac{1}{2\pi}\int_0^{2\pi}\log[2l/a\sin(\phi/2)]d\phi \\ &\quad + (a^2/8\pi l^2)\int_0^{2\pi}\sin^2(\phi/2)d\phi + \cdots \\ &= \log(4l/a) + a^2/8l^2 + \cdots. \quad (23)\end{aligned}$$

The error is still of the order of a^2/l^2 and Hallén's approximation of the double integral by the simple integral is justified for many practical antennas.

Thus, Brillouin's suggestion that Hallén's formula for the input impedance yields poor results on account of the approximations in (1) is not borne out by the above analysis. On the other hand, we are now in a position to call attention to several points which, so far, seem to have been overlooked. Equation (14) is an identity only so long as $I(z)$ is not infinite. The integral equation itself gives no indication with regard to finiteness of $I(z)$; but from other considerations it is known that for an infinitesimal gap, $I(0)$ is infinite. Hence, infinitesimal gaps must certainly be excluded from consideration if we are to use (14). Furthermore, the first term on the right of (14) cannot be a good approximation to the integral on the left if $I(\xi)$ varies rapidly at $\xi = z$; this is precisely the case in the vicinity of $z = 0$ when $s \ll a$. Hence, transformation (14) requires s to be sufficiently large to insure that $I(z)$ does not vary too rapidly.

Finally, the first term on the right of (14) is a better approximation to the integral on the left if βr is small; that is, if $2\pi l/\lambda$ is small. It will be recalled that $\Omega(z)$ as given by (15) is proportional to the direct-current inductance per unit length of a uniform current filament extending from $z = -l$ to $z = l$. If we multiply (5) by $i\omega\mu$, we should find, by comparing with Maxwell's equations, that the right-hand side represents the dynamic component of the voltage per unit length along the antenna. Hallén's first approximation, $I_0(z)$ to the current is obtained in effect by dividing this voltage by $i\omega L$, where L is the direct-current inductance of the wire. It would have been better to use the "alternating-current inductance." Mathematically this would correspond to the following transformation:

$$\int_0^{2\pi}\int_{-l}^{l}\frac{I(\xi)e^{-i\beta r}}{2\pi r}\,d\xi d\phi = I(z)\int_0^{2\pi}\int_{-l}^{l}\frac{\cos\beta r}{2\pi r}\,d\xi d\phi + \int_0^{2\pi}\int_{-l}^{l}\frac{I(\xi)-I(z)}{2\pi r}\cos\beta r\,d\xi d\phi - i\int_0^{2\pi}\int_{-l}^{l}\frac{I(\xi)}{2\pi r}\sin\beta r\,d\xi d\phi \qquad (24)$$

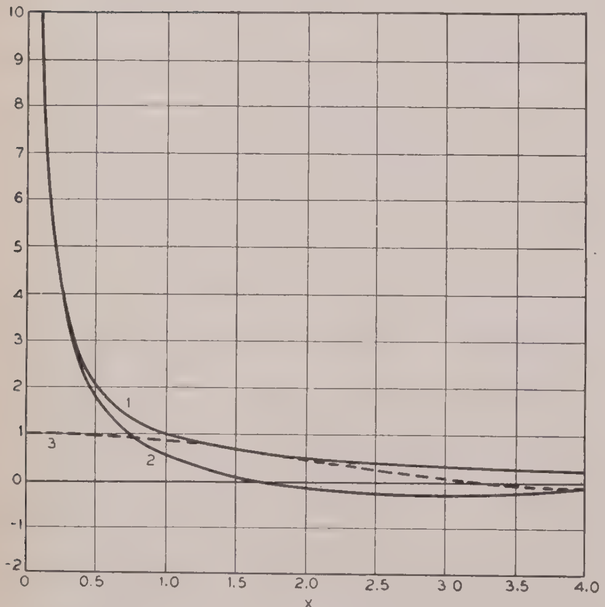

Fig. 3—Curves 1, 2, and 3 represent respectively $1/x$, $\cos x/x$ and $\sin x/x$.

in place of (14). This is precisely the transformation employed by Gray. Instead of $\Omega(z)$ we now have

$$P(z) = \int_0^{2\pi}\int_{-l}^{l}\frac{\cos\beta r}{2\pi r}\,d\xi d\phi \simeq 2\log(\lambda/\pi a) - 2C + \mathrm{Ci}\,\beta(l+z) + \mathrm{Ci}\,\beta(l-z), \qquad (25)$$

where $C=0.577\cdots$ is Euler's constant. At $z=0$, we have

$$P(0) = 2\log(\lambda/\pi a) - 2C + 2\,\mathrm{Ci}\,\beta l. \qquad (26)$$

With the exception of this modification, there is no reason why the process of successive approximation should not be carried on as suggested by Hallén. In (17) we replace Ω by P and r by $|\xi-z|$. Gray improves the first correction term by using *under the integral sign* the average value of $P(z)$, rather than $P(0)$; but this is not obligatory, and one could use $P(0)$ throughout, relying on the higher-order terms for greater accuracy.

Another possible transformation, preferable to (14), is

$$\int_0^{2\pi}\int_{-l}^{l}\frac{I(\xi)e^{-i\beta r}}{2\pi r}\,d\xi d\phi = I(z)\int_0^{2\pi}\int_{-l}^{l}\frac{e^{-i\beta r}}{2\pi r}\,d\xi d\phi + \int_0^{2\pi}\int_{-l}^{l}\frac{[I(\xi)-I(z)]}{2\pi r}e^{-i\beta r}d\xi d\phi. \qquad (27)$$

The function corresponding to $P(z)$ is now

$$\widehat{P}(z) = P(z) - i[\mathrm{Si}\,\beta(l+z) + \mathrm{Si}\,\beta(l-z)]. \qquad (28)$$

Fig. 3 illustrates the reason why (14) does not yield as good a series for $I(z)$ as (24), particularly when $\lambda/4<l<\lambda/2$. The real part of the integrand contains the factor $\cos\beta r/\beta r$, which vanishes when $r=\lambda/4$. When $l=\lambda/2$, this factor vanishes almost where $I(\xi)$ is maximum; hence, the region in the vicinity of $r=\lambda/4$ contributes little to the real part of the integral, and this feature is reflected in the first term on the right-hand side of (24). On the other hand, the corresponding region contributes a substantial amount to the first term on the right-hand side of (14). This difference is particularly important because the region $r<0.03\lambda$, where $\cos\beta r/\beta r$ is large and nearly equal to $1/\beta r$, contributes little to the integral in question on account of smallness of $I(\xi)$. Fig. 3 also shows that (27) will not be as good as (24) as the basis for successive approximations. Since $\cos\beta r/r$ is large only when $\xi=z$, the real part of the integral on the left is nearly proportional to the current $I(z)$ at $\xi=z$; on the other hand, $\sin\beta r/r$ is of the same order of magnitude in a large interval about $\xi=z$; and the imaginary part is not even approximately proportional to $I(z)$.

AN INFINITELY LONG ANTENNA

Infinitely long antennas are of interest to us primarily because they can be treated in at least three quite different ways, and a comparison will be instructive. In order to approximate an infinitely long antenna in a laboratory, one should make it several wavelengths long and attempt to terminate so as to produce substantially progressive waves for several wavelengths on both sides of the generator.

If the radius is small compared with $\lambda/4$, the simplest method is the one which takes advantage of the fact that in an infinitely long antenna we have to consider only the principal progressive waves.[9] Thus when z/a is large, we have

$$I(z) = Ae^{-i\beta z}/\sqrt{K(z)},\quad K(z) = 120\log(2z/a), \qquad (29)$$

where A is a constant and $K(z)$ is the nominal characteristic impedance. The input impedance is found to be approximately[10]

$$Z_i = 120\left(\log\frac{\lambda}{2\pi a} - C - i\,\mathrm{Si}\,2\pi\right) = 120\log(\lambda/2a) - 207 - i172. \qquad (30)$$

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[9] S. A. Schelkunoff, "Principal and complementary waves in antennas," to be published in PROC. I.R.E., January, 1946.

[10] J. Stratton and L. J. Chu, "Steady-state solutions of electromagnetic field problems, I. Forced oscillations of a cylindrical conductor," *Jour. Appl. Phys.*, vol. 12, pp. 230–235; March, 1941.

From conservation of energy we obtain

$$A^2 = R_i[I(0)]^2, \tag{31}$$

where R_i is the input resistance; therefore,

$$I(z) = \frac{I(0)\sqrt{R_i}e^{-i\beta z}}{\sqrt{K(z)}} \simeq \frac{Ve^{-i\beta z}}{\sqrt{R_iK(z)}}, \tag{32}$$

except for a possible constant phase shift, unobtainable from the energy equation.

Several years ago, we tried to solve this problem for any radius. It is relatively easy to express the current in the following form:

$$I(z) = \frac{V(0)a}{30\lambda s}\int_{(C)} \frac{\sinh(\gamma s/2)K_1(a\sqrt{-\gamma^2-\beta^2})e^{\gamma z}}{\gamma\sqrt{-\gamma^2-\beta^2}\,K_0(a\sqrt{-\gamma^2-\beta^2})}d\gamma \tag{33}$$

where s is the length of the gap and (C) is the imaginary axis indented to the right of the branch point $\gamma=-i\beta$ and to the left of $\gamma=i\beta$. The admittance is then $I(s/2)/V(0)$. The chief difficulty is the evaluation of the integral. If a is small, we succeeded in approximating the real part of the admittance[2]

$$G_1 = \frac{1}{120M}\left[1 + \frac{\log 2}{2M} - \frac{\frac{2}{3}\pi^2 - (\log 2)^2}{4M^2} + \cdots\right],$$

$$M = \log(\lambda/2\pi a) + \tfrac{1}{2}\log 2 - C,\ C = 0.577\cdots. \tag{34}$$

It is to be noted that, if we retain only the first two terms of this equation, the reciprocal of G_i is equal to the real part of Z_i in (30). When a is large, the real part of the integral can be evaluated numerically; by this method a previous curve[11] was obtained.

Finally when a is small, we were successful in obtaining from (33) the following approximation:

$$I(z) = \frac{V(0)}{120[\log(\lambda/2\pi a) - C]} \cdot\left[1 - \frac{\log(2\pi z/\lambda) + \log 2 + C}{2[\log(\lambda/2\pi a) - C]}\right]e^{-i\beta z}, \tag{35}$$

except for a term in quadrature with $V(0)$. The approximation assumes that z/a is large but z/λ is small enough to make the second term in brackets fairly small. This equation agrees with (32), as may be verified by using

$$\begin{aligned} K(z) &= 120\log(2z/a) \\ &= 120\left[\left(\log\frac{\lambda}{2\pi a} - C\right) + \left(\log\frac{2\pi z}{\lambda} + \log 2 + C\right)\right] \end{aligned} \tag{36}$$

and expanding $\sqrt{K(z)}$ in a power series. Equation (32), however, is good for *all* large values of z/a.

All these results are at variance with those obtained[10] by Stratton and Chu, from an equation which is equivalent to (33) with $s=0$. These authors find that the current in an infinitely long, perfectly conducting cylinder vanishes and that nonzero current exists only in imperfect wires. The discrepancy is explained by the incomplete evaluation of the integral; they evaluate only the contributions from the poles of the integrand, and for perfectly conducting wires the latter has no poles.

Let us now consider the integral equation for an infinitely long wire. The exact equation is

$$60\int_{-\infty}^{\infty}\int_0^{2\pi}\frac{I(\xi)e^{-i\beta r}}{2\pi r}d\phi d\xi = Ve^{-i\beta|z|},\quad |z| > s/2;$$

$$r = \sqrt{(\xi - z)^2 + \rho^2},\quad \rho = 2a\sin(\phi/2). \tag{37}$$

The approximate equation, when a is small, is

$$60\int_{-\infty}^{\infty}\frac{I(\xi)e^{-i\beta r}}{r}d\xi = Ve^{-i\beta|z|},\quad |z| > s/2;$$

$$r = \sqrt{(\xi - z)^2 + a^2}. \tag{38}$$

Identity (14) breaks down, this time because $\Omega=2\log(2l/a)$ becomes infinite. On the other hand, (24) is still valid and we write

$$\begin{aligned} I(z) &= \frac{Ve^{-i\beta|z|}}{K} + \frac{60}{K}\int_{-\infty}^{\infty}\frac{I(z) - I(\xi)}{|\xi - z|}\cos\beta(\xi - z)d\xi \\ &\quad + \frac{i60}{K}\int_{-\infty}^{\infty}\frac{\sin\beta(\xi - z)}{\xi - z}I(\xi)d\xi, \end{aligned} \tag{39}$$

where

$$K = 60\int_{-\infty}^{\infty}\frac{\cos\beta r}{r}d\xi = 120\left(\log\frac{\lambda}{\pi a} - C\right). \tag{40}$$

In the first approximation we have

$$I(z) = (V/K)e^{-i\beta|z|}. \tag{41}$$

Substituting this in (39), we have the second approximation for positive values of z

$$\begin{aligned} I(z) = (V/K)e^{-i\beta z}\Bigg[1 &- \frac{60}{K}\left(\log\beta z + C - \log 2 - \frac{i\pi}{2}\right) \\ &- \frac{60}{K}\left(-\operatorname{Ci}2\beta z + i\operatorname{Si}2\beta z - \frac{i\pi}{2}\right)e^{2i\beta z}\Bigg]. \end{aligned} \tag{42}$$

When z is nearly zero, we have

$$I(0) = \frac{V}{K}\left[1 + \frac{120\log 2}{K} + i\frac{60\pi}{K}\right]. \tag{43}$$

From this we obtain the input impedance

$$\begin{aligned} Z_i &= K - 120\log 2 - i60\pi \\ &= 120\left(\log\frac{\lambda}{2\pi a} - C\right) - i60\pi. \end{aligned} \tag{44}$$

Comparing this to (30), we find the real parts in agreement; the imaginary parts are 172 and 188 ohms, respectively.

If βz is fairly large, then (42) becomes

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[11] See Fig. 22 of footnote reference 2.

$$I(z) = (V/K)e^{-i\beta z}\left[1 - \frac{60}{K}\left(\log\frac{2\pi z}{\lambda} + C - \log 2 - \frac{i\pi}{2}\right)\right]. \quad (45)$$

This equation checks (32) and (35), except for a phase shift which was not included in the previous equations.

CONCLUSION

Our general conclusion is that there is no disagreement between the antenna theories developed directly from Maxwell's equations and Hallén's theory based on the integral equation *provided the latter is treated with sufficient care.* On the surface, the solution of the integral equation seems to require only elementary algebra and rudiments of integral calculus, but appearances are sometimes deceptive. Just to give substance to the warning, let us return to the exact equation (37) for the infinitely long cylinder. Among its solutions is to be found the internal wave as well as the external; that is, the well-known guided wave inside a hollow-cylindrical shell. Once this solution has been determined by other methods, it is easy to verify that it satisfies (37); but it would be easy to overlook this particular solution, let alone obtain it.

Similarly, the equation for natural oscillations on a closed, perfectly conducting surface is satisfied by the *internal undamped* oscillations as well as by the external damped oscillations. This is true, for instance, in the case of Brillouin's equation for natural oscillations on a cylinder. We have in mind the equation which is obtained when we equate to zero the tangential electric intensity calculated by the retarded potential method from the current distribution on the surface. This equation is not the equation for the natural oscillations on a perfectly conducting *solid* or in a cavity imbedded in a perfectly conducting solid. For the solid, the restrictions on the electric intensity are more stringent since it must vanish everywhere within the solid.

Thus we have shown that the approximations involved in Hallén's integral equation are justified, so long as the square of the ratio of the length of the antenna to its diameter is large compared with the natural logarithm of this ratio; but his method of successive approximations to the solution and his impedance formula are based on a tacit initial assumption that the antenna is short compared with the wavelength, and due to the nature of his series, the higher-order approximations fail to correct this initial assumption even for thinnest practicable antennas. This method has been relied upon in several recent papers on antennas, and the conclusions contained in them would be similarly affected. On the other hand, the solution of Hallén's equation as given by Gray does not involve the above-mentioned stringent assumption, and for this reason is more satisfactory. Her method can thus be recommended as an alternative to the method presented in our previous paper.[2]

Frequency-Modulation Distortion Caused by Multipath Transmission*

MURLAN S. CORRINGTON†

Summary—**When a frequency-modulated wave is received by more than one path, so that two or more of the voltages which are induced in the antenna have nearly the same amplitude, considerable distortion can result. Large objects, such as hills or high buildings, can reflect and absorb the waves and thus cause interference. Two interfering waves will be in phase part of the time and out of phase part of the time during modulation. This causes amplitude modulation on the resultant carrier, and a sharp irregularity in the instantaneous frequency, corresponding to each hole in the carrier caused by the interference.**

The distortion can often be greatly reduced by detuning the receiver or discriminator until the hole in the carrier coincides with the zero-balance point of the discriminator. Directional antennas are helpful when the signals from the desired station are not coming from the same direction.

Formulas are derived for the modulated envelope and for the distortion. A Fourier-series analysis of the distorted audio output makes it possible to calculate the effects of de-emphasis networks and other audio selectivity.

INTRODUCTION

DURING a demonstration of a frequency-modulation receiver in New York in the spring of 1942, difficulty was experienced in locating an antenna in the RCA Building which would produce a signal of quality comparable with that known to be transmitted by station W2XWG. The antenna was located in a room in this steel building on the side away from the transmitter. The distortion was present even on relatively small deviations.

The tests were repeated in an apartment house on 93rd Street, and similar distortion was encountered. Further tests were made later with a second frequency-modulation reciver to determine whether the distortion

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was caused by a defective receiver or was introduced during the transmission of the signal. Both receivers produced the same type of distorted output. Since intermediate-frequency amplifier selectivity would not cause this effect, it was suggested that it might be caused by multipath reception. During July and August of 1942, further observations on a qualitative basis were made on the eighty-fifth floor of the Empire State Building. Path differences due to reflections were encountered which were of sufficient magnitude to cause partial or complete cancellation of the waves at one or more frequencies in a 150-kilocycle-wide channel as used by frequency-modulation broadcast stations. A large number of field tests have been made since that time, and they show that this trouble can occur in nearly any location in Manhattan as well as in many of the surrounding towns in northern New Jersey.

In midtown Manhattan there are many steel buildings which cause multipath transmission. If the receiver is inside such a structure, an outside antenna is desirable. If such an antenna is not practicable, it probably will be necessary to use more than one inside receiving antenna in order to obtain full performance on the high-fidelity programs from all the New York frequency-modulation stations. This will be true even if the receiver has the best performance characteristics now commercially available. For a single stationary antenna, distortion will probably be noticed in many locations on at least one of the stations.

Cases have been observed where the movement of a person around in the room near the receiving antenna has changed the relative field strengths enough so that, when the person was in some parts of the room, the reception was satisfactory, and when he was in other parts if was unsatisfactory. A parasitic element such as a metal rod or dipole can sometimes be moved around in a room to produce the same effect. Maximum distortion occurs at the time of cancellation of the two waves. This causes a great reduction in the voltage in the intermediate-frequency channel and it may fall below the threshold of limiting in the receiver.

If the cancellation occurs near one edge of the band, the signal may sound fairly good during the soft passages but it becomes distorted on the loud passages. It will then sound like an overloaded audio-amplifier stage. If the cancellation occurs near the middle of the channel, the entire output may be so distorted that it becomes almost unintelligible; it then sounds somewhat like severe selective fading in amplitude modulation. Since high-order harmonics are produced, the distortion may appear as buzzes, rattles, or swishes, and if it occurs during the playing of a phonograph record, one may think the pickup is not tracking properly. The high frequencies often tend to become irritating due to intermodulation.

Two amateurs[1] have reported distortion of the programs from WMIT when the receiver is in the mountainous region near Asheville, N. C. At times they found that the audio signal was badly distorted to the point of becoming unintelligible, even for strong signals, and that the distortion is a constant phenomenon over periods of weeks in those spots where it occurs. DuMont and Goldsmith[2] found that multipath conditions are common in which two signals of approximately equal strength arrive at a receiving antenna, and therefore this type of distortion with frequency-modulation sound transmission can be expected to occur frequently.

A similar type of distortion had previously been encountered in long-distance transmission. In 1930, Eckersley[3] was working with amplitude-modulation transmitters which had incidental frequency modulation. Because of this frequency shift he found "most appalling distortion" resulting from delayed echoes caused by reflections from the Heaviside layer. He found that it was necessary to use special precautions to keep any frequency shift from getting into the amplitude-modulation transmitters. The distortion when frequency modulation and amplitude modulation were both present was worse than that for amplitude modulation only.

A study of frequency-modulation propagation over long distances was made by Crosby.[4] A transmitter was set up in Bolinas, California, and the transmissions were observed on receivers at the Riverhead, New York, station. He found that there was considerable distortion, and that in some cases a signal which gave fair intelligibility on amplitude modulation was practically unintelligible on frequency modulation. He stated, "The general conclusion derived from the tests and theory is that, on circuits where multipath transmission is encountered, frequency modulation is impracticable." In a later paper,[5] he again reported similar distortion. Other studies[6,7] have shown that reflections from the ground and nearby buildings cause multipath transmission, for both horizontal and vertical polarization, which can introduce distortion. Reflections from airplanes flying overhead sometimes cause interference for short intervals.

The same difficulty is encountered in television reception. The higher frequencies used tend to increase the difficulty, since the phase changes encountered are greater. This causes light and dark bands in the picture and results in synchronization difficulties when frequency modulation is used on video or synchronizing signals.

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[1] A. D. Mayo and Charles W. Sumner, "F. M. distortion in mountainous terrain," *QST*, vol. 28, pp. 34–36; March, 1944.

[2] Allen B. DuMont and Thomas T. Goldsmith, Jr., "Television broadcast coverage," Proc. I.R.E., vol. 32, pp. 192–205; April, 1944.

[3] T. L. Eckersley, "Frequency modulation and distortion," *Exp. Wireless and the Wireless Eng.*, vol. 7, pp. 482–484; September, 1930.

[4] Murray G. Crosby, "Frequency-modulation propagation characteristics," Proc. I.R.E., vol. 24, pp. 898–913; June, 1936.

[5] Murray G. Crosby, "Observations of frequency-modulation propagation on 26 megacycles," Proc. I.R.E., vol. 29, pp. 398–403; July, 1941.

[6] P. S. Carter and G. S. Wickizer, "Ultra-high-frequency transmission between the RCA building and the Empire State Building in New York City," Proc. I.R.E., vol. 24, pp. 1082–1094; August, 1936.

[7] R. W. George, "A study of ultra-high-frequency wide-band propagation characteristics," Proc. I.R.E., vol. 27, pp. 28–35; January, 1939.

Since field tests have demonstrated the possibility of encountering distortion due to multipath propagation even though the distance from the transmitter to the receiver is relatively short, a theoretical and experimental study was made to determine the factors contributing to this type of distortion. It is the purpose of this paper to present the results of this investigation.

THEORETICAL CONSIDERATIONS

Suppose that one frequency-modulated wave is delayed with respect to the second by a time interval t_0 because of two-path transmission. The equations for the instantaneous voltages of the two waves are

$$e_1 = E_1 \sin \left\{ \omega t + \frac{D}{\mu} \sin 2\pi\mu t \right\} \tag{1}$$

$$e_2 = E_2 \sin \left\{ \omega(t - t_0) + \frac{D}{\mu} \sin 2\pi\mu(t - t_0) \right\} \tag{2}$$

where ω = unmodulated-carrier angular frequency
D = maximum frequency deviation
μ = audio frequency
t_0 = time delay of the second wave with respect to the first
E_1 = amplitude of first wave
E_2 = amplitude of second wave.

By combining these two waves by the parallelogram law, the equation for the resultant voltage becomes, as shown in Appendix I, equation (22),

$$e_1 + e_2 = E_1\sqrt{1 + x^2 + 2x \cos \left\{ z \cos (2\pi\mu t - \pi\mu t_0) + \omega t_0 \right\}} \\ \sin \left[\omega t + \frac{D}{\mu} \sin 2\pi\mu t - \tan^{-1} \frac{x \sin \left\{ z \cos (2\pi\mu t - \pi\mu t_0) + \omega t_0 \right\}}{1 + x \cos \left\{ z \cos (2\pi\mu t - \pi\mu t_0) + \omega t_0 \right\}} \right] \tag{3}$$

where

$$z = 2 \frac{D}{\mu} \sin \pi\mu t_0 \tag{4}$$

and

$$x = \frac{E_2}{E_1}. \tag{5}$$

The first part of this expression (that under the radical sign) gives the variations in amplitude of the resultant carrier, and the argument of the sine function shows the variations in phase of the signal. There is considerable amplitude modulation introduced, and the frequency-modulated signal is also distorted. The resulting carrier envelope will be considered now, and the distortion in the audio output will be discussed later.

Effect of Interference on the Envelope of the Radio-Frequency Wave

The resultant amplitude of the radio-frequency carrier is given by

$$|e_1+e_2| = E_1\sqrt{1+x^2+2x \cos \left\{ z \cos (2\pi\mu t - \pi\mu t_0) + \omega t_0 \right\}}. \tag{6}$$

The amplitude is a function of the ratio of the two signal voltages x, the maximum deviation D, the time delay t_0, and the audio frequency μ. The only effect of the carrier frequency ω is to determine the initial phase angle ωt_0.

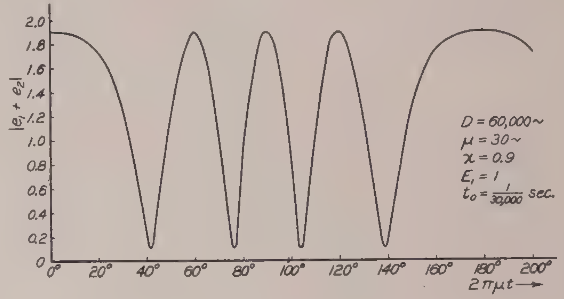


Fig. 1—Carrier envelope.

Fig. 1 shows the carrier envelope for a deviation of 60,000 cycles per second from a mean carrier frequency of 45 megacycles per second. E_1 was assumed to be one volt and E_2 was taken as 0.9 volt, so the voltages are nearly equal. The conditions were chosen so that the two radio-frequency signal voltages were in phase at the undeviated position corresponding to $2\pi\mu t = 90$ degrees. The equation of the envelope becomes

$$|e_1+e_2| = \sqrt{1.81+1.80 \cos \left\{ 4\pi \cos (2\pi\mu t - 0.18°) \right\}}. \tag{7}$$

The path difference corresponding to t_0 is 6.2 miles. Zero degrees correspond to full deviation on one side.

Since the audio signal is a cosine function, it varies slowly at first, then more rapidly as it goes through 90 degrees, and then slows down as it approaches 180 degrees. Fig. 1 shows how the spacing of the peaks and dips of the carrier envelope corresponds to this variation. Zero degrees corresponds to the slowest rate of change, and the peak is very wide; at 90 degrees, the variation is most rapid, and the peak is narrow. The amplitude can be represented by the resultant of two rotating vectors $\boldsymbol{R}$ as shown by Fig. 2. If $\boldsymbol{E_2}$ were to rotate at a uniform rate, the resultant $\boldsymbol{R}$ (which corresponds to the instantaneous value of the envelope), would go through identical cycles, and the peaks of Fig. 1 would be evenly spaced. When the carrier wave is modulated sinusoidally, however, the vector $\boldsymbol{E}_2$ does not rotate uniformly, but rocks back and forth, with a sinusoidal variation of the angle θ. Under the conditions represented by Fig. 1, the vector $\boldsymbol{E}_2$ starts with $\theta = 0.18$ degrees and makes two revolutions (4π radians) while the carrier frequency varies from maximum deviation to the mean undeviated frequency. Each time it goes past point B it gives a hole in the carrier envelope, and when it goes through A it gives a peak.

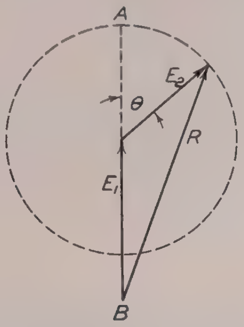


Fig. 2—Resultant of two signals.

Fig. 3 shows what happens when the frequency is increased from 30 cycles per second to 5000 cycles per second. The equation of this envelope is

$$|e_1+e_2| = \sqrt{1.81+1.80\cos\{12\cos(2\pi\mu t-30°).\}} \qquad (8)$$

Since $\pi\mu t_0 = \pi/6$ radians $= 30$ degrees, the curve is shifted over approximately 30 degrees from that in Fig. 1, corresponding to one half the fraction of the audio cycle by which one wave is lagging behind the other.

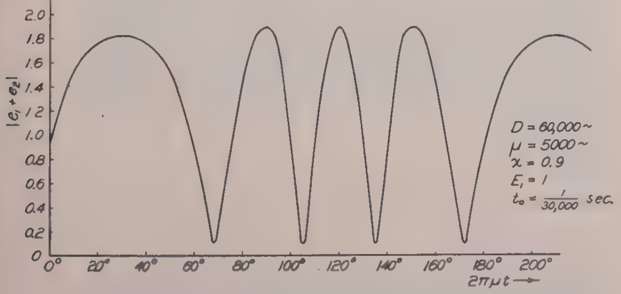


Fig. 3—Carrier envelope.

Another interesting difference is that the two voltages come almost, but not quite, into phase at 30 degrees. This peak is therefore slightly lower than the others. The difference is caused by the fact that the sine of an angle is not linear, but departs from a straight line for larger angles. The maximum angle through which the one vector of Fig. 2 rotates with respect to the other is given by

$$z = 2\frac{D}{\mu}\sin\pi\mu t_0 \text{ radians} \qquad (9)$$

and this sine function introduces the nonlinearity. Further changes of the frequency merely move the curve along the axis, but effect slight changes in shape.

Variation of the Envelope with Deviation

Fig. 4 shows the effect of changes of the maximum deviation D. The constants were chosen such that the two voltages are always in phase when $2\pi\mu t = 0$ degrees. For a small deviation, say $D = 15{,}000$ cycles per second, the two voltages start to go out of phase as $2\pi\mu t$ increases, but do not go completely out of phase. The maximum angle between the vectors in this case is

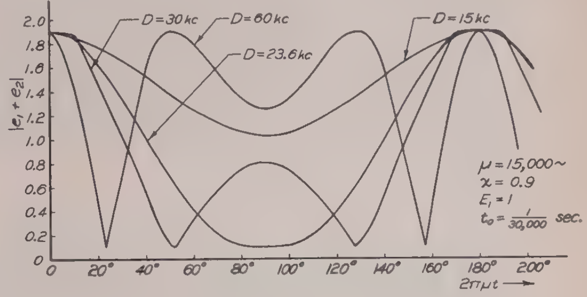


Fig. 4—Variation of carrier envelope with D.

$$\begin{aligned} z &= 2\frac{D}{\mu}\sin\pi\mu t_0 \\ &= 2\frac{15{,}000}{15{,}000}\sin\frac{\pi(15{,}000)}{30{,}000} \\ &= 2 \text{ radians.} \end{aligned}$$

The vector $\boldsymbol{E}_2$ rotates clockwise 2 radians, then reverses and goes 2 radians counterclockwise, as shown by Fig. 5. This process gives the envelope shown in Fig. 4.

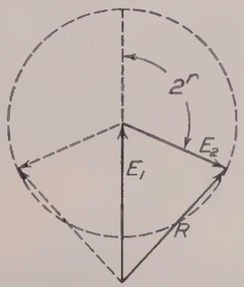
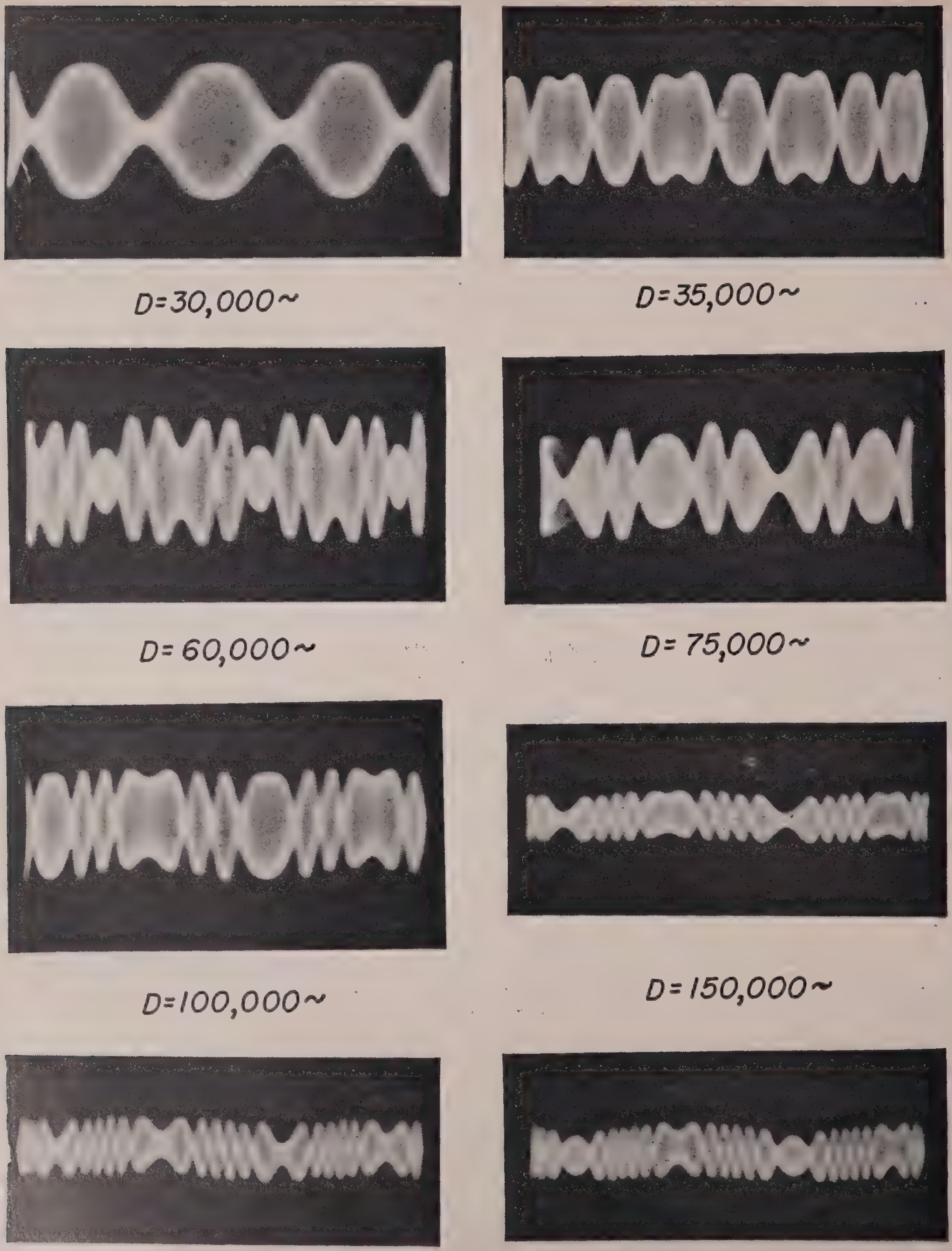


Fig. 5—Variation of resultant.

If the maximum deviation D is increased to 23,600 cycles per second, the two voltages go just out of phase in the mean position. A further increase in D causes them to start to come into phase again, as shown by the curve for $D = 30{,}000$ cycles per second. Further increases in D increase the number of peaks and holes in the envelope.

Variation of the Envelope with Path Difference

A change in the path difference for the two waves will also change the number of peaks and holes. Fig. 6 shows the envelopes for different path differences from 1.0 mile to 6.2 miles. The curves are very similar to those of



D=30,000~

D=35,000~

D=60,000~

D=75,000~

D=100,000~

D=150,000~

D=175,000~

D=200,000~

Fig. 9—Oscillograms of carrier envelopes.

Fig. 4. The shapes can be changed further by adjusting the initial phase of the two waves.

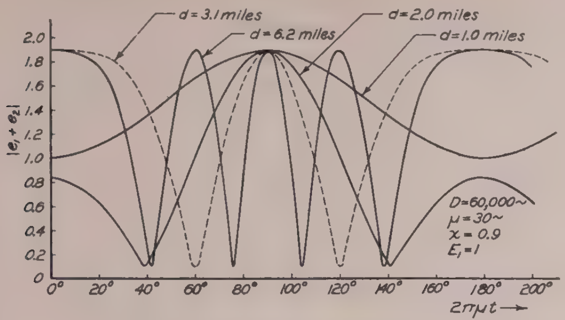


Fig. 6—Variation of carrier envelope with path difference.

Variation of the Amplitude with Deviation.

When the wave is modulated sinusoidally, the variation of the frequency deviation is not linear, and this causes the unequal spacing of the peaks and dips as shown in Fig. 1. If the same set of conditions is used, and the variation of amplitude is plotted against the deviation, the spacing of the peaks and dips will be uniform as shown by Fig. 7. A further increase in deviation will cause more peaks, but will not alter the spacing if the time delay is unchanged.

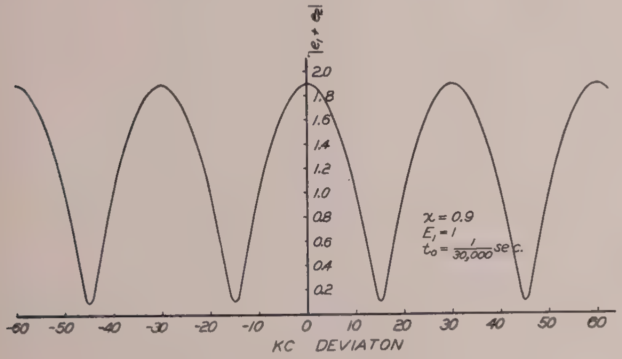


Fig. 7—Variation of amplitude with deviation.

The equation of the amplitude is

$$|e_1 + e_2| = \sqrt{1.81 + 1.80 \cos\left(\frac{2\pi D}{30{,}000}\right)} \tag{10}$$

if the waves start out in phase.

Experimental Study of Carrier Envelope

An artificial time delay was obtained by passing the signal through 10,500 feet of high-frequency cable, and then combining it with a signal from the same source which had traveled only a short distance. The circuit is shown in Fig. 8.

The frequency-modulation signal generator was modulated by a sinusoidal audio tone, and the potentiometer P was adjusted until the two voltages on the oscilloscope were nearly equal. The maximum deviation D was adjustable, and the initial phase could be adjusted by shifting slightly the mean carrier frequency. A series of oscillograms was taken that showed the carriers beating together for different deviations D. It took approximately 30,000 cycles per second deviation to cause the carrier to go from in-phase to out-of-phase.

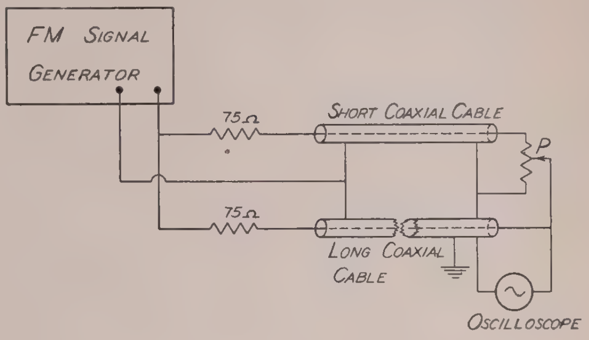


Fig. 8—Circuit for studying interference.

The oscillograms of Fig. 9 show how the number of peaks and holes varies with the deviation D. They are seen to be very similar to those of Figs. 1, 3, 4, and 6. In each one, the two voltages were nearly equal.

Fig. 10 shows an oscillogram in which one signal is about 3.5 times as strong as the second. The general shape is the same as those of Fig. 9, but the holes do not go nearly so deep. The velocity of propagation in the cable was 66 per cent of that in free space; therefore, the effective path difference was 15,900 feet. This corresponds to a time delay of 16.2 microseconds.

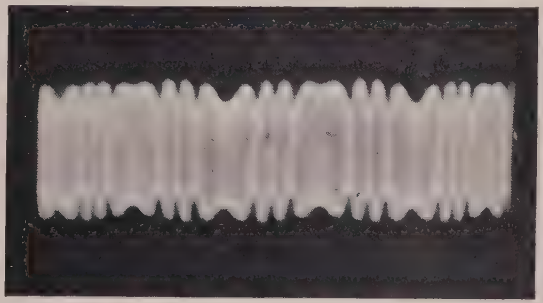


Fig. 10—Oscillogram of carrier envelope.

Determination of Time Delay

Let the path difference be d and the time delay corresponding to this difference be t_0. The unmodulated carrier frequency is ω. Let λ_1 be the wavelength corresponding to ω, and let λ_2 equal the wavelength corresponding to $\omega+D$, where D is the frequency deviation required to increase by one the number of wavelengths in d.

Then

$$d = N\lambda_1 = (N+1)\lambda_2 \tag{11}$$

and

$$d = N\frac{3 \times 10^8}{\omega} = (N+1)\frac{3 \times 10^8}{\omega + D} \text{ meters} \tag{12}$$

where $\omega\lambda_1 = 3 \times 10^8$ meters per second, the free-space velocity of the wave, and $(\omega+D)\lambda_2=3\times10^8$ meters per second. Eliminate N from the second equation

$$d=\frac{d\omega}{\omega+D}+\frac{3\times10^8}{\omega+D}$$

and

$$d=\frac{3\times10^8}{D}\text{ meters.} \tag{13}$$

The time delay

$$t_0=\frac{d}{3\times10^8}=\frac{1}{D}\text{ seconds.} \tag{14}$$

If D is the frequency shift required to go from out-of-phase, to in-phase, and out-of-phase again, the time delay will be $1/D$.

Field Tests

To study the distortion encountered in New York City, the NBC transmitter at the Empire State Building was operated with 300-cycle pure-tone modulation and a deviation of ±75 kilocycles. A medium-priced receiver and an antenna were set up in a room next to the transmitter. The antenna could easily be located to give two-path reception, with a time delay of about seven microseconds. Figs. 11 and 12 show the carrier envelope and the discriminator output. In Fig. 11 the two waves canceled each other near one end of the swing. The resultant amplitude modulation which passed the limiter, and the coincidental phase modulation gave the result shown in Fig. 12. One half of the cycle is relatively free from distortion, since the two signals did not go out of phase at any part of that swing. The two irregularities in the audio output correspond to the point of cancellation. This distortion could be eliminated by correct antenna location.

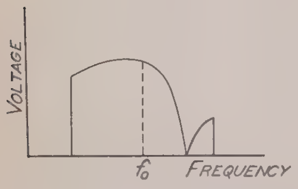


Fig. 11—Carrier envelope.

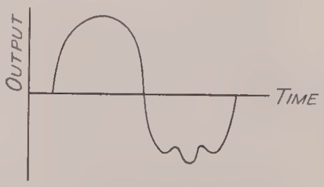


Fig. 12—Discriminator output.

In the RCA Building on the side away from the transmitter, the antenna was fairly well shielded from the direct wave from the transmitter, and it received signals which may have been reflected from other buildings. Fig. 13 shows several curves of the audio output and the corresponding variation in the carrier voltage during the cycle. They are all for a 300-cycle audio tone and ±75-kilocycle deviation, for various antenna positions. The audio output should be a pure sine wave, and the intermediate-frequency voltage should show the intermediate-frequency selectivity curve, as shown by the dotted lines. Although there were several irregularities along the straighter portions, the most serious distortion occurs at the ends of the swing. The first intermediate-frequency voltage curve shows that there were several paths of nearly equal voltage. In the second case, the secondary paths were still present, but the relative voltage was much less, as shown by the decreased departure from the selectivity curve. Since the signal strength was fairly low, some of the distortion was probably due to the inability of the limiter to remove all the deep holes in the carrier.

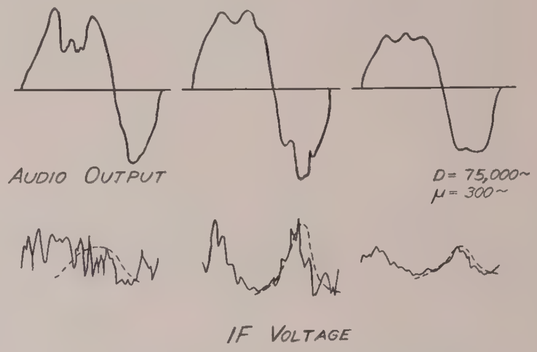


Fig. 13—Distortion for various antenna positions.

Many other field tests in New York City and northern New Jersey have shown this distortion. The time delay is usually about 7 to 12 microseconds. This means that, for two-path transmission, there will be one or two holes in the carrier. Since the limiter tends to fail at these holes, the distortion can often be reduced by detuning the receiver until the hole in the carrier coincides with the discriminator zero-balance frequency. Since the discriminator output is zero at this point, variations in input voltage cause a minimum of distortion. It sometimes happens, when the direct and indirect waves are nearly equal, that severe distortion is encountered in the audio output of the set even though the input signal, as indicated by the voltage developed at the limiter grid, would normally be sufficient to secure limiting.

Effect of Interference on the Audio Output from the Discriminator

As shown in Appendix I, equation (23), the audio output from the discriminator is proportional to

$$D\cos 2\pi\mu t+\frac{2D\sin\pi\mu t_0\sin(2\pi\mu t-\pi\mu t_0)}{\dfrac{1/x+\cos\{z\cos(2\pi\mu t-\pi\mu t_0)+\omega t_0\}}{x+\cos\{z\cos(2\pi\mu t-\pi\mu t_0)+\omega t_0\}}+1}. \tag{15}$$

The first term represents the undistorted output, and the second term shows the distortion introduced by the interference. This result was obtained by assuming that a limiter removes the coincidental amplitude modulation from the wave, and that the discriminator output is proportional to the instantaneous frequency.

General Case—Two Voltages Nearly Equal

The distortion term can be expanded in a Fourier series, as shown in Appendix II. The audio output from the discriminator is proportional to

$$\text{output}=D\cos 2\pi\mu t$$

$$-2\mu\sum_{n=0}^{\infty}(-1)^n C(2n+1, z; x, \omega t_0)(2n+1)\sin(2n+1)\gamma$$

$$-2\mu\sum_{n=1}^{\infty}(-1)^n S(2n, z; x, \omega t_0)(2n)\sin 2n\gamma \qquad (16)$$

where

$\gamma = 2\pi\mu t - \pi\mu t_0$ and the C and S functions are defined as follows:

$$C(m, n; x, \theta) = \sum_{s=1}^{\infty}\frac{(-x)^s}{s}J_m(sn)\cos s\theta. \qquad (17)$$

$$S(m, n; x, \theta) = \sum_{s=1}^{\infty}\frac{(-x)^s}{s}J_m(sn)\sin s\theta. \qquad (18)$$

The amplitude of each harmonic can be calculated from these equations by assuming the proper value of n. The C and S functions can be computed from tables of Bessel functions. Figs. 14, 15, 16, and 17 show some graphs of these functions.

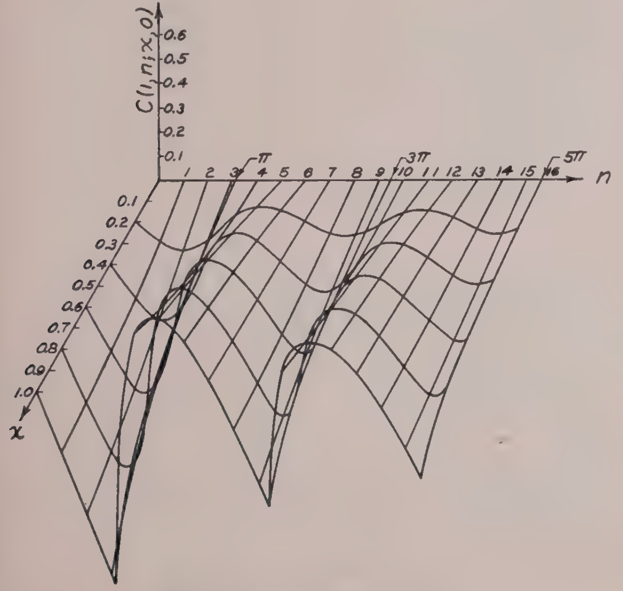


Fig. 14—Graph of $C(1, n; x, 0)$.

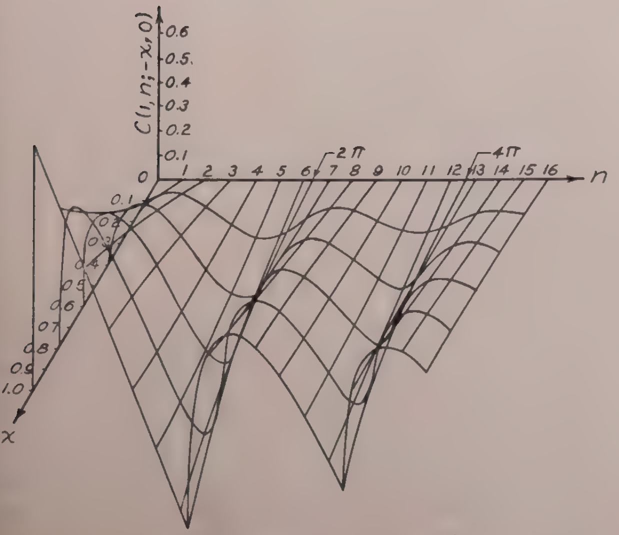


Fig. 15—Graph of $C(1, n; -x, 0)$.

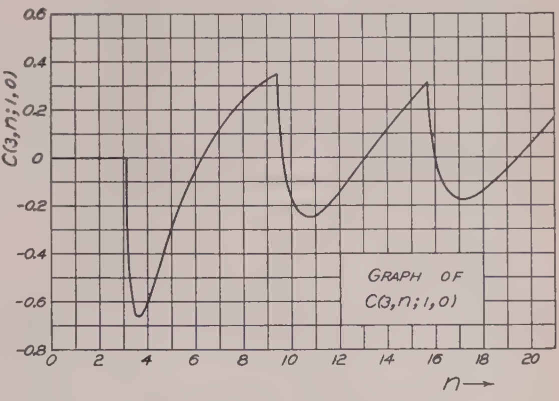


Fig. 16—Graph of $C(3, n; 1, 0)$.

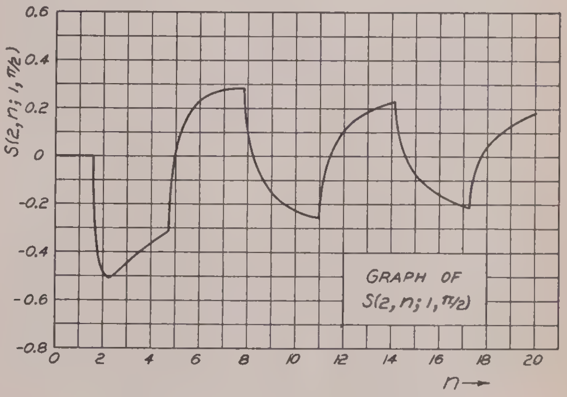


Fig. 17—Graph of $S(2, n; 1, \pi/2)$.

Fig. 18 shows the distortion caused at an audio frequency of 500 cycles per second under the conditions listed. The phase of the two radio-frequency waves was chosen such that they are 180 degrees out of phase at the unmodulated frequency. This gives the maximum of distortion. The first irregularity comes at about 10 degrees, where the vectors are still rotating relatively slowly with respect to each other. By the time the angle $2\pi\mu t = 63$ degrees, the two vectors are rotating rapidly and a sharp dip is created. A similar effect occurs at 93 degrees and

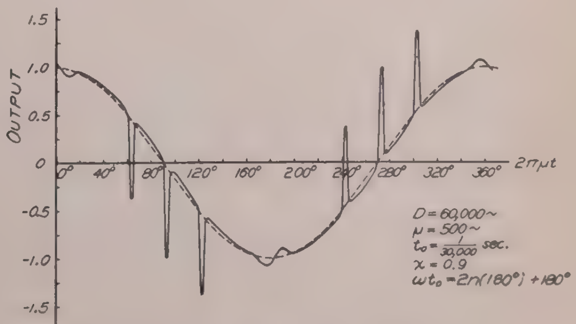


Fig. 18—Distorted audio output.

123 degrees. At 180 degrees the vectors are again rotating slowly, and the peak is small. The process is repeated during the second half of the audio cycle.

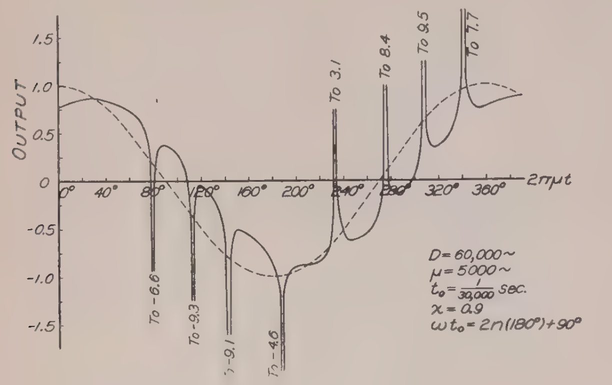


Fig. 19—Distorted audio output.

Fig. 19 shows the distorted audio output when the audio frequency is increased to 5000 cycles per second. The value for ωt_0 was chosen to make the two radio-frequency signals start 90 degrees out of phase.

Fig. 20 shows the two vectors $\boldsymbol{E}_1$ and $\boldsymbol{E}_2$ and the angle θ, which is the angle between them. The resultant $\boldsymbol{R}$ makes an angle α with $\boldsymbol{E}_1$. The amount the distorted curve of Fig. 19 departs from the pure cosine wave (shown by the dotted line) is proportional to the time rate of change of angle α. It is easy to see from Fig. 20 that, as the angle θ nears and goes through 180 degrees, there is a very great change in α for a relatively small change in θ, when $\boldsymbol{E}_1$ and $\boldsymbol{E}_2$ are nearly the same length. This means that the first derivative of α with respect to time is very large, and this causes the deep dips and high peaks. The nearer $\boldsymbol{E}_2$ comes to $\boldsymbol{E}_1$ in absolute magnitude, the greater the departure from the undistorted curve. In the limit, as $\boldsymbol{E}_1 = \boldsymbol{E}_2$, $\alpha = +90$ degrees as θ approaches 180 degrees from the right and $\alpha = -90$ degrees as θ passes through 180 degrees. This means a 180-degree change in α as θ passes 180 degrees. The derivative of α with respect to time is therefore infinite at that instant, and the curve representing the audio output becomes a sharp pulse.

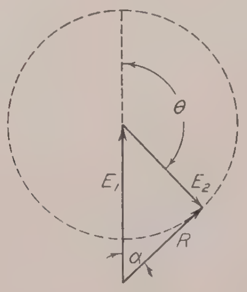


Fig. 20—Variation of α.

This theory assumes an infinite bandwidth and linear phase-shift in the receiver. The very deep and narrow dips and peaks show the presence of harmonics of high order which are in phase at each peak. Since the receiver has a limited bandwidth, many of these harmonics will be filtered out in the audio system, so the audio output will not be actually as distorted as shown. The effect of nonlinear phase shift and limited bandwidth is to broaden out and shorten the peaks in the output. The undistorted cosine curve is shown for comparison. The envelope of the carrier for this set of conditions is shown by Fig. 21. There is one dip or peak corresponding to each hole in the envelope of the carrier.

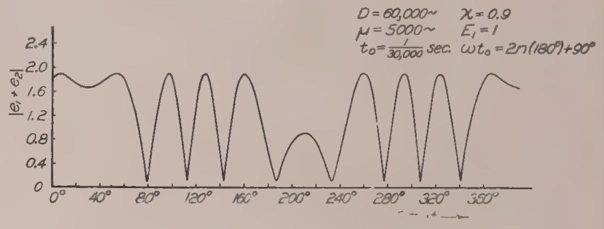


Fig. 21—Carrier envelope.

If the maximum deviation is reduced, such that the two vectors do not go out of phase so often, the distortion will be much less, because the number of peaks will be reduced and the one vector will not rotate so fast relative to the second. Fig. 22 shows the output at 1000 cycles per second with a maximum deviation of 15,000 cycles per second. The time delay is 1/30,000 second as before. The phase of the two radio-frequency carriers was chosen such that they were in phase at the undeviated frequency. This assumption also reduces the distortion. Upon reducing the deviation further, or decreasing the time delay of the retarded signal, the irregularities will disappear.

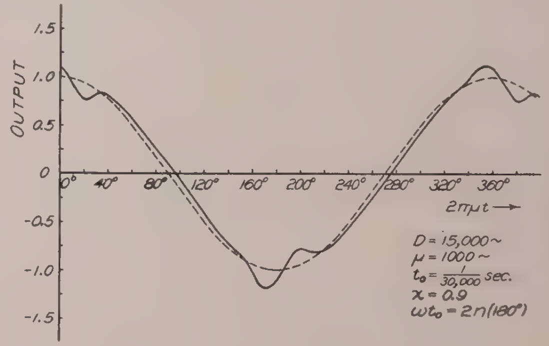
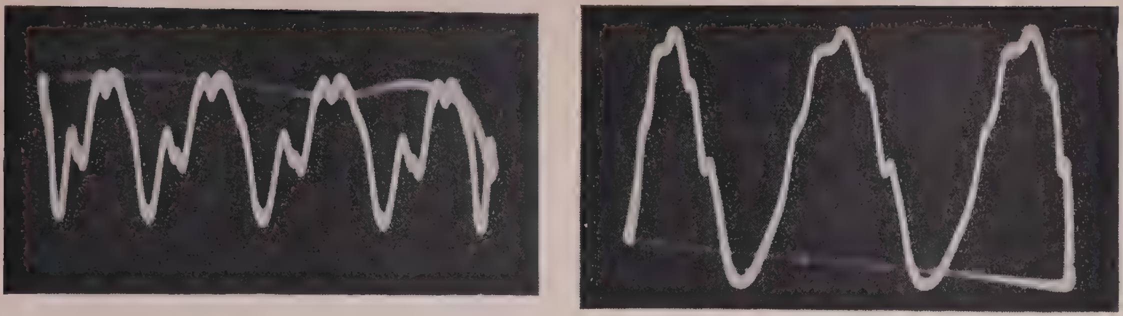


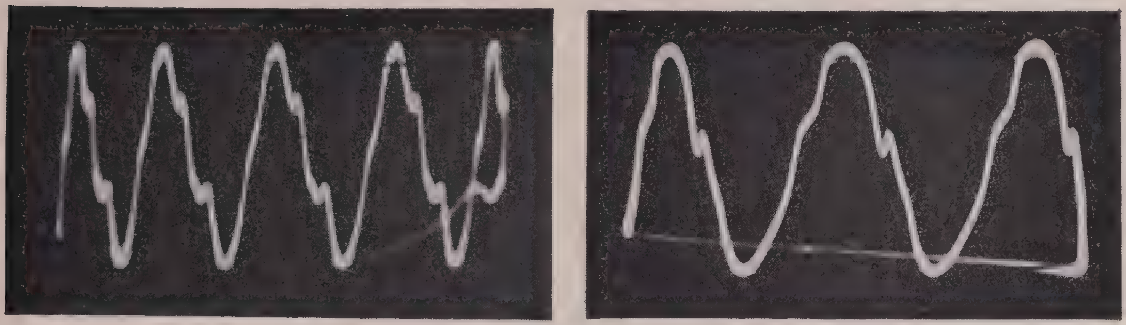
Fig. 22—Distorted audio output.

Experimental Study of Distorted Output

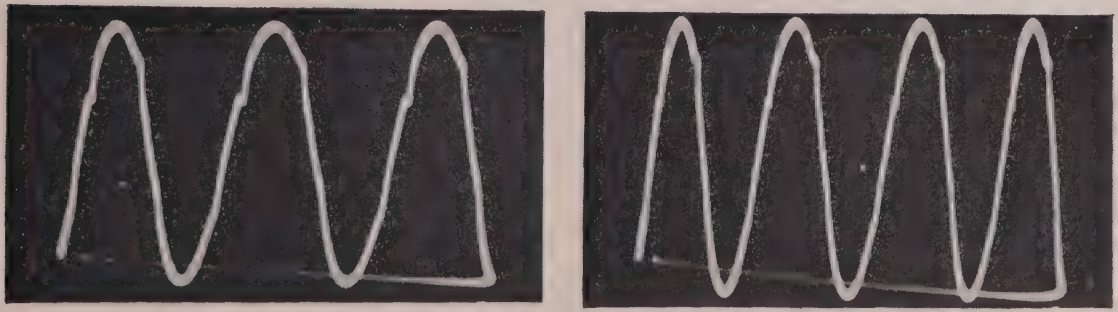
The distortion in the audio output from the receiver was studied by using the circuit of Fig. 8 with a frequency-modulation receiver connected just ahead of the oscilloscope. The signal was passed through the intermediate-frequency amplifier, limiter, and discriminator, and the distortion was studied with the oscilloscope. Fig. 23 shows a series of oscillograms taken with different audio frequencies and deviations.



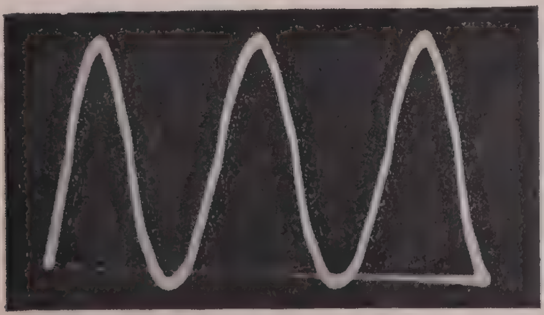
$D = 100{,}000$~ $\mu = 10{,}000$~ $D = 75{,}000$~ $\mu = 10{,}000$~



$D = 100{,}000$~ $\mu = 4000$~ $D = 75{,}000$~ $\mu = 4000$~



$D = 100{,}000$~ $\mu = 1000$~ $D = 75{,}000$~ $\mu = 1000$~



$D = 75{,}000$~ $\mu = 400$~

Fig. 23—Oscillograms showing distortion.

It is easily seen that high deviations and high audio frequencies cause the most distortion. Reducing either one will decrease the distortion. These curves cannot be compared directly with the theoretical ones previously shown, since the time delay is not the same. The time delay used experimentally was limited to 16.2 microseconds by the available length of cable, while the theoretical curves were based on a time delay of 33.3 microseconds. Naturally, the increased time delay causes the one signal to lag further behind and thus causes greater distortion. Crosby[8] has also shown many oscillograms of this type of distortion, taken over a path from Kansas City to Riverhead, Long Island. These show the characteristic sharp breaks in the audio output and the effect of noise.

CONCLUSION

The calculations and experimental study show that serious distortion of a frequency-modulation signal can be caused by multipath transmission over relatively short distances. This distortion is liable to be encountered in the vicinity of large buildings or other objects which reflect and absorb the waves and thus cause interference. These buildings or other large objects can be near either the transmitter or the receiver and cause such reflections. The two signals will be in phase part of the time and out of phase part of the time during the audio cycle. If the signals have approximately equal amplitudes, nearly complete cancellation results when they go out of phase. This change in amplitude modulates the resultant carrier since it causes a sudden drop or deep hole in the resultant carrier amplitude, as shown by Figs. 1, 3, 9, etc. If these holes are so deep that the limiter in the conventional receiver cannot hold the output constant, amplitude modulation will be impressed on the discriminator. If the discriminator can respond to these variations in amplitude, the output will be distorted at the part of the cycle which corresponds to the hole in the resultant carrier. If the resultant-carrier amplitude drops below the noise level, the distorted part of the cycle will have noise superimposed on it. Cases have been observed in which the output was undistorted at all parts of the cycle except at one point, and there the output was all noise.

Each time the two nearly-equal carriers go out of phase, there is a very rapid phase shift which causes a peak or dip in the output corresponding to this phase modulation. The worst distortion occurs when the two signals are of nearly equal voltage, and when the two radio-frequency carriers are out of phase when the modulation is removed. The distortion increases with the audio frequency because the higher the audio frequency, the further the retarded wave lags behind, in terms of an audio cycle. An increase in the maximum frequency deviation increases the number of times the two signals go in and out of phase in each audio cycle, and this also increases the distortion. This increase of the distortion as the deviation is increased is in contrast to the reduction of ordinary noise as the bandwidth of the transmission is increased.

The amplitude of each harmonic can be calculated by the formulas given. From these values, the effect of a low-pass audio filter can be determined. The low frequencies are not distorted very much, and the high frequencies will have their distortion reduced by the audio selectivity. This means that the medium frequencies will be the ones which are distorted the most and they are the ones most needed for intelligibility. For shorter path differences the time delay is less, and the distortion is reduced. If the two signals are approximately in phase at the undeviated position, they cannot go entirely out of phase for path differences less than about one mile with the deviations used today. However, if they are out of phase at the undeviated position, distortion will be produced even for small time delays. For small deviations and low audio frequencies the signals do not get far out of phase; thus there is not much distortion.

The distortion due to multipath transmission of frequency-modulated waves contains many high-order harmonics and is, therefore, different from amplitude-modulation distortion caused by the same process. In amplitude modulation the harmonics are all of low order and are not so disagreeable. Frequency-modulation distortion often sounds like crackles, rattles, swishes, or gurgles and may not be objectionable on low modulation. It sometimes makes one wonder if the transmitter is being operated properly. At times it is noticeable only on the loud passages and may sound somewhat like overloading in an audio-amplifier tube. At other times the distortion may be so bad that the signal is almost unintelligible; it then sounds somewhat like selective fading in amplitude modulation.

It is doubtful whether an increase in selectivity can decrease the distortion. The hole in the resultant carrier and the coincidental phase modulation can occur at any part of the audio cycle, and thus can occur at the carrier frequency as well as any other. This cancellation at the center frequency cannot be eliminated by greater selectivity. An increase in sensitivity usually helps to decrease the distortion, since it helps to maintain sufficient carrier amplitude to operate the limiter at all times. A commercial frequency-modulation receiver designed to reduce multipath distortion should have a sensitivity of at least 10 or 20 microvolts, and for that signal strength should remove all amplitude variations in the signal even at the ends of the swing. The receiver should also remove the holes from the carrier; otherwise, distortion may be noted on some stations that may have much higher signal strengths than others which have no apparent distortion.

A properly oriented directional antenna, such as a dipole, will reduce the distortion if the two signals are coming from different directions, as is likely to be true near large buildings. Such antennas are of limited value when the signals are coming from the same direction. A power-line antenna or built-in loop often gives multipath trouble, and little can be done to reduce it since most such antennas are not ordinarily adjustable, and the voltage pickup is usually less than for a dipole.

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[8] See page 401 of footnote reference 5.

In some receiving locations, the antenna should preferably be remote from the receiver so the relative field strengths will not be changed appreciably as the occupants of the room move about. Cases have been observed where the signal was either distorted or undistorted depending upon where one stood in the room. For time delays of approximately 6 or 8 microseconds, the distortion can often be greatly reduced by mistuning the receiver or discriminator until the hole in the carrier coincides with the zero-balance point of the discriminator, since any amplitude modulation that passes the limiter then causes very little audio output. However, this may cause an increase in noise. Under these conditions, a tuning meter is of little value.

APPENDIX I

DERIVATION OF EQUATIONS FOR FREQUENCY-MODULATION DISTORTION CAUSED BY MULTIPATH TRANSMISSION

The analysis of this appendix is similar to that developed by Crosby,[4] and is repeated here for convenience. It will serve as an introduction to the derivation in Appendix II.

Let one frequency-modulated wave be delayed with respect to another by a given time interval t_0. This condition can be caused by a two-path transmission, where one path is longer than the other by a fixed distance. The equations for the instantaneous voltages of the two signals are

$$e_1 = E_1 \sin\left(\omega t + \frac{D}{\mu}\sin 2\pi\mu t\right) \tag{19}$$

$$e_2 = E_2 \sin\left\{\omega(t - t_0) + \frac{D}{\mu}\sin 2\pi\mu(t - t_0)\right\} \tag{20}$$

where

ω = unmodulated-carrier angular frequency
D = maximum frequency deviation
μ = audio frequency
t_0 = time delay of the second signal with respect to the first.

These two waves can be combined by the parallelogram law. In Fig. 24, the law of cosines gives

$$R = \sqrt{E_1^2 + E_2^2 + 2E_1E_2\cos\theta}.$$

The angle θ between the two vectors E_1 and E_2 equals the difference of the two arguments of the sine functions.

$$\theta = \frac{D}{\mu}\sin 2\pi\mu t - \frac{D}{\mu}\sin 2\pi\mu(t - t_0) + \omega t_0.$$

The angle α between R and E_1 is given by

$$\tan\alpha = \frac{E_2\sin\theta}{E_1 + E_2\cos\theta} = \frac{x\sin\theta}{1 + x\cos\theta}$$

where $$x = \frac{E_2}{E_1}.$$

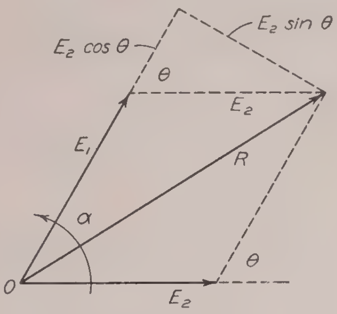

Fig. 24—Composition of two vectors.

The law of cosines gives, therefore

$$e_1 + e_2 = \sqrt{E_1^2 + E_2^2 + 2E_1E_2\cos\theta}$$
$$\sin\left(\omega t + \frac{D}{\mu}\sin 2\pi\mu t - \tan^{-1}\frac{x\sin\theta}{1 + x\cos\theta}\right). \tag{21}$$

Since $\sin\alpha - \sin(\alpha - \beta) = +2\cos(\alpha - \beta/2)\sin\beta/2$,

$$\theta = 2\frac{D}{\mu}\sin\pi\mu t_0\cos(2\pi\mu t - \pi\mu t_0) + \omega t_0$$
$$= z\cos(2\pi\mu t - \pi\mu t_0) + \omega t_0$$

where $z = 2D/\mu \sin\pi\mu t_0$.

The equation for the resultant voltage becomes

$$e_1 + e_2$$
$$= \sqrt{E_1^2 + E_2^2 + 2E_1E_2\cos\{z\cos(2\pi\mu t - \pi\mu t_0) + \omega t_0\}}$$
$$\sin\left[\omega t + \frac{D}{\mu}\sin 2\pi\mu t\right.$$
$$\left. - \tan^{-1}\frac{x\sin\{z\cos(2\pi\mu t - \pi\mu t_0) + \omega t_0\}}{1 + x\cos\{z\cos(2\pi\mu t - \pi\mu t_0) + \omega t_0\}}\right]. \tag{22}$$

The amplitude term of this expression shows the variation of the resultant carrier amplitude, and the argument of the sine function gives the frequency modulation and resulting distortion. When the signal goes through a perfect limiter the amplitude term is suppressed, and the argument of the sine term is effective at the discriminator.

Calculation of Audio Output

The output from the discriminator is proportional to the instantaneous frequency, where the instantaneous frequency[9] is defined by

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[9] J. R. Carson, "Notes on the theory of modulation," PROC. I.R.E., vol. 10, pp. 57–64; February, 1922.

$$f = \frac{1}{2\pi}\frac{d}{dt}(\text{argument of sine function})$$

$$= \frac{1}{2\pi}\frac{d}{dt}\left[\omega t + \frac{D}{\mu}\sin 2\pi\mu t - \tan^{-1}\frac{x\sin\{z\cos(2\pi\mu t - \pi\mu t_0)+\omega t_0\}}{1 + x\cos\{z\cos(2\pi\mu t - \pi\mu t_0)+\omega t_0\}}\right].$$

Since

$$\frac{d}{dx}\tan^{-1}u = \frac{1}{1+u^2}\frac{du}{dx}$$

and

$$\frac{d}{dx}\left(\frac{u}{v}\right) = \frac{v\frac{du}{dx} - u\frac{dv}{dx}}{v^2}$$

the derivative can be simplified to give

$$f = \frac{\omega}{2\pi} + D\cos 2\pi\mu t + \frac{2D\sin\pi\mu t_0\sin(2\pi\mu t - \pi\mu t_0)}{\dfrac{1/x + \cos\{z\cos(2\pi\mu t-\pi\mu t_0)+\omega t_0\}}{x+\cos\{z\cos(2\pi\mu t - \pi\mu t_0)+\omega t_0\}} + 1}. \quad (23)$$

A balanced discriminator is operated so there is no output at the unmodulated carrier frequency ω, and the output is proportional to the deviation from this frequency. The audio output is therefore proportional to the second and third terms of (23). The third term represents the distortion, and it can be serious.

APPENDIX II

FOURIER SERIES ANALYSIS OF DISTORTED AUDIO OUTPUT

The third term of (23) of Appendix I represents the distortion in the audio output from the discriminator. It was obtained by differentiation of the argument of the sine function of (22). Consider this term

$$\frac{d}{dt}\tan^{-1}\frac{x\sin\{z\cos(2\pi\mu t-\pi\mu t_0)+\omega t_0\}}{1+x\cos\{z\cos(2\pi\mu t - \pi\mu t_0)+\omega t_0\}} = \frac{d}{dt}\tan^{-1}\frac{x\sin\theta}{1+x\cos\theta} \quad (24)$$

where $\theta = z\cos(2\pi\mu t - \pi\mu t_0)+\omega t_0$.

Let

$$\alpha = \tan^{-1}\frac{x\sin\theta}{1+x\cos\theta} \quad (25)$$

so

$$\tan\alpha = \frac{x\sin\theta}{1+x\cos\theta}.$$

From Fig. 25

$$k\sin\alpha = x\sin\theta \quad (26)$$

$$k\cos\alpha = 1 + x\cos\theta \quad (27)$$

where

$$k = \sqrt{(1+x\cos\theta)^2 + (x\sin\theta)^2}$$

$$= \sqrt{1 + x^2 + 2x\cos\theta}. \quad (28)$$

Multiply (26) by i and add (27).

$$1 + x\cos\theta + ix\sin\theta = k(\cos\alpha + i\sin\alpha).$$

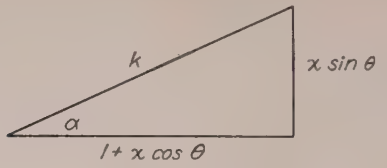

Fig. 25—Determination of k.

Change these complex numbers to the exponential form,

$$1 + xe^{i\theta} = ke^{i\alpha} \quad (29)$$

and take logarithms of both sides,

$$\log(1 + xe^{i\theta}) = \log k + i\alpha. \quad (30)$$

Expand this in a power series by the well-known relation

$$\log(1+x) = x - \frac{x^2}{2} + \frac{x^3}{3} - \frac{x^4}{4} + \cdots \quad -1 < x \leqq 1 \quad (31)$$

so

$$\log(1+xe^{i\theta}) = xe^{i\theta} - \frac{x^2}{2}e^{2i\theta} + \frac{x^3}{3}e^{3i\theta} - \frac{x^4}{4}e^{4i\theta} + \cdots$$

$$= x(\cos\theta + i\sin\theta) - \frac{x^2}{2}(\cos 2\theta + i\sin 2\theta)$$

$$+ \frac{x^3}{3}(\cos 3\theta + i\sin 3\theta)$$

$$- \frac{x^4}{4}(\cos 4\theta + i\sin 4\theta) + \cdots. \quad (32)$$

Equate the imaginary terms of (32), and use (30).

$$\alpha = x\sin\theta - \frac{x^2}{2}\sin 2\theta + \frac{x^3}{3}\sin 3\theta - \frac{x^4}{4}\sin 4\theta + \cdots. \quad (33)$$

Differentiate this expression with respect to time,

$$\frac{d\alpha}{dt} = (x\cos\theta - x^2\cos 2\theta + x^3\cos 3\theta - x^4\cos 4\theta + \cdots)\frac{d\theta}{dt}$$

$$= 2\pi\mu z\sum_{n=1}^{\infty}(-x)^n\sin(2\pi\mu t - \pi\mu t_0)$$

$$\cos\{nz\cos(2\pi\mu t - \pi\mu t_0) + n\omega t_0\}. \quad (34)$$

Two lemmas will now be proved.

Lemma α

$$e^{iz\cos\theta} = \sum_{k=-\infty}^{\infty} i^k J_k(x)e^{ik\theta}.$$

Two expansions in series of Bessel coefficients due to Jacobi[10] are

$$\cos(x\cos\theta) = J_0(x) + 2\sum_{n=1}^{\infty}(-1)^n J_{2n}(x)\cos 2n\theta$$

and $\sin(x\cos\theta) = 2\sum_{n=0}^{\infty}(-1)^n J_{2n+1}(x)\cos(2n+1)\theta.$

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[10] G. N. Watson, "A treatise on the theory of Bessel functions," The Macmillan Company, New York, N. Y., Second edition, 1944, p. 22.

Multiply the second equation by i and add the first equation. Change the cosine terms to the complex form. This gives

$$\cos(x\cos\theta)+i\sin(x\cos\theta)$$

$$=J_0(x)+\sum_{n=1}^{\infty}(-1)^nJ_{2n}(x)e^{2in\theta}+\sum_{n=1}^{\infty}(-1)^nJ_{2n}(x)e^{-2in\theta}$$

$$+i\sum_{n=0}^{\infty}(-1)^nJ_{2n+1}(x)e^{(2n+1)i\theta}$$

$$+i\sum_{n=0}^{\infty}(-1)^nJ_{2n+1}(x)e^{-(2n+1)i\theta}.$$

Since $(-1)^n=i^{2n}=i^{-2n}$ and $J_n(x)=(-1)^nJ_{-n}(x)$, this proves the lemma.

Lemma β

$$\sin\theta\cos(a\cos\theta+b)$$

$$=2\cos b\sum_{m=0}^{\infty}(-1)^m\frac{2m+1}{a}J_{2m+1}(a)\sin(2m+1)\theta$$

$$+2\sin b\sum_{m=1}^{\infty}(-1)^m\frac{2m}{a}J_{2m}(a)\sin 2m\theta.$$

In complex form, using lemma α,

$$\sin\theta\cos(a\cos\theta+b)$$

$$=\frac{1}{4i}\left\{e^{i\theta}-e^{-i\theta}\right\}\left\{e^{i(a\cos\theta+b)}+e^{-i(a\cos\theta+b)}\right\}$$

$$=\frac{1}{4i}\left\{e^{i(\theta+b)}\sum_{k=-\infty}^{\infty}i^kJ_k(a)e^{ik\theta}+e^{i(\theta-b)}\sum_{k=-\infty}^{\infty}(-i)^kJ_k(a)e^{ik\theta}\right.$$

$$\left.-e^{-i(\theta-b)}\sum_{k=-\infty}^{\infty}i^kJ_k(a)e^{ik\theta}-e^{-i(\theta+b)}\sum_{k=-\infty}^{\infty}(-i)^kJ_k(a)e^{ik\theta}\right\}$$

$$=\frac{1}{4i}\left\{\sum_{k=-\infty}^{\infty}i^kJ_k(a)\left[e^{i(k+1)\theta+ib}-e^{i(k-1)\theta+ib}\right]\right.$$

$$\left.+\sum_{k=-\infty}^{\infty}(-i)^kJ_k(a)\left[e^{i(k+1)\theta-ib}-e^{i(k-1)\theta-ib}\right]\right\}$$

$$=\frac{1}{4i}\sum_{k=-\infty}^{\infty}\left\{i^{k-1}J_{k-1}(a)e^{i(k\theta+b)}-i^{k+1}J_{k+1}(a)e^{i(k\theta+b)}\right.$$

$$\left.+(-i)^{k-1}J_{k-1}(a)e^{i(k\theta-b)}-(-i)^{k+1}J_{k+1}(a)e^{i(k\theta-b)}\right\}$$

$$=-\frac{1}{4}\sum_{k=-\infty}^{\infty}\left\{J_{k-1}(a)+J_{k+1}(a)\right\}\left\{i^ke^{i(k\theta+b)}\right.$$

$$\left.-(-i)^ke^{i(k\theta-b)}\right\}$$

$$=-\frac{1}{2}\sum_{k=-\infty}^{\infty}i^k\frac{k}{a}J_k(a)\left\{e^{i(k\theta+b)}-(-1)^ke^{i(k\theta-b)}\right\}$$

$$=-\frac{1}{2}\sum_{k=1}^{\infty}\frac{k}{a}i^kJ_k(a)\left\{e^{i(k\theta+b)}+(-1)^ke^{-i(k\theta+b)}\right.$$

$$\left.-(-1)^ke^{i(k\theta-b)}-e^{-i(k\theta-b)}\right\}$$

$$=\sum_{m=0}^{\infty}(-1)^m\frac{2m+1}{a}J_{2m+1}(a)\left\{\sin\left[(2m+1)\theta+b\right]\right.$$

$$\left.+\sin\left[(2m+1)\theta-b\right]\right\}$$

$$-\sum_{m=1}^{\infty}(-1)^m\frac{2m}{a}J_{2m}(a)\left\{\cos(2m\theta+b)-\cos(2m\theta-b)\right\}.$$

The identities

$$\sin(x+y)+\sin(x-y)=2\sin x\cos y$$

and

$$\cos(x+y)-\cos(x-y)=-2\sin x\sin y$$

prove the lemma.

An application of lemma β to (34) gives the relation

$$\frac{d\alpha}{dt}=2\pi\mu z\sum_{n=1}^{\infty}(-x)^n$$

$$\left\{\cos n\omega t_0\sum_{m=0}^{\infty}(-1)^m\frac{2(2m+1)}{nz}J_{2m+1}(nz)\sin(2m+1)\gamma\right.$$

$$\left.+\sin n\omega t_0\sum_{m=1}^{\infty}(-1)^m\frac{2(2m)}{nz}J_{2m}(nz)\sin 2m\gamma\right\}$$

$$=2\pi\mu\sum_{n=1}^{\infty}\sum_{m=0}^{\infty}(-1)^m(-x)^n$$

$$\frac{\cos n\omega t_0}{n}2(2m+1)J_{2m+1}(nz)\sin(2m+1)\gamma$$

$$+2\pi\mu\sum_{n=1}^{\infty}\sum_{m=1}^{\infty}(-1)^m(-x)^n$$

$$\frac{\sin n\omega t_0}{n}2(2m)J_{2m}(nz)\sin 2m\gamma \tag{35}$$

where $\gamma=2\pi\mu t-\pi\mu t_0$. The audio output from the discriminator is therefore proportional to

$$\text{output}\propto D\cos 2\pi\mu t$$

$$-2\mu\sum_{n=0}^{\infty}(2n+1)(-1)^nC(2n+1,z;x,\omega t_0)\sin(2n+1)\gamma$$

$$-2\mu\sum_{n=1}^{\infty}(2n)(-1)^nS(2n,z;x,\omega t_0)\sin 2n\gamma \tag{36}$$

where the C and S functions are defined as follows:

$$C(m,n;x,\theta)=\sum_{s=1}^{\infty}\frac{(-x)^s}{s}J_m(sn)\cos s\theta$$

$$S(m,n;x,\theta)=\sum_{s=1}^{\infty}\frac{(-x)^s}{s}J_m(sn)\sin s\theta.$$

The graphs of Figs. 14, 15, 16, and 17 show how these functions vary for the particular values of the parameters which were chosen. Other graphs of the functions can be plotted by using tables of Bessel functions[11] which were prepared for this purpose.

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[11] Murlan S. Corrington and William Miehle, "Tables of Bessel functions $J_n(x)$ for large arguments," *Jour. Math. and Phys.* (M.I.T.), vol. 24, pp. 30–50; February, 1945.

Symmetrical Antenna Arrays*

CHARLES W. HARRISON, JR.†, MEMBER, I.R.E.

Summary—A relatively simple method is presented for calculating the impedance properties of antenna arrays consisting of n identical radiators oriented at the vertices of regular polygons. All antennas are required to carry currents of equal magnitude, but not necessarily of the same phase. However, the choice of phase angle must be such that the required electrical symmetry of the array is unimpaired.

The case of a circular array with a central radiator is discussed.

Immediate practical applications of the theory include:

(a) Determination of the driving-point impedance of the primary radiator in a corner-reflector antenna.

(b) Calculation of the power lost through radiation, or conversely the signal pickup, of certain multiple-wire transmission lines, when operated in a resonant condition.

(c) Analyses of problems involving symmetrically disposed fixed antennas for direction-finding purposes.

(d) Ascertaining the effectiveness of circular arrays for transmission.

Several numerical illustrations of the theory are given.

GENERAL THEORY OF SYMMETRICAL ANTENNA ARRAYS

CONSIDER an array consisting of five identical antennas symmetrically disposed on the circumference of a circle, as shown in Fig. 1. Each antenna is of half-length h and radius a. The origin for co-ordinates ($z=0$) is the meridian plane of the array. Let it be assumed that the antennas are driven by identical generators maintaining the flow of equal currents. These currents may differ in phase by an appropriate angle. By direct analogy with antenna theory previously formulated,[1–5] the distribution of current along antenna (1), for example, is given by

$$I_z = -\frac{j4\pi}{\Omega R_c}\left\{C_1 \cos \beta z + \tfrac{1}{2}V_o^e \sin \beta |z|\right\}$$

$$-\frac{1}{\Omega}\left\{I_z \ln\left(1-\frac{z^2}{h^2}\right) + I_z\delta + \int_{-h}^{+h} \frac{I_z' e^{-j\beta r_{11}} - I_z}{r_{11}}\, dz'\right\}$$
$$-\frac{1}{\Omega}\left\{e^{-j\theta_{12}}\int_{-h}^{+h} I_z' \frac{e^{-j\beta r_{12}}}{r_{12}}\, dz' + e^{-j\theta_{13}}\int_{-h}^{+h} I_z' \frac{e^{-j\beta r_{13}}}{r_{13}}\, dz'\right.$$
$$+ e^{-j\theta_{14}}\int_{-h}^{+h} I_z' \frac{e^{-j\beta r_{14}}}{r_{14}}\, dz'$$
$$\left.+ e^{-j\theta_{15}}\int_{-h}^{+h} I_z' \frac{e^{-j\beta r_{15}}}{r_{15}}\, dz'\right\}. \tag{1}$$

The following notation is used:

$$\Omega = 2 \ln \frac{2h}{a}. \tag{2}$$

R_c is the intrinsic resistance of free space.

$$R_c = 376.7 \approx 120\pi \text{ ohms}. \tag{3}$$

C_1 is an arbitrary constant, to be evaluated from the boundary condition

$$I_h = 0, \qquad (z = \pm h).$$

V_o^e is the voltage applied across the input terminals of any radiator.

$$\beta = \frac{2\pi}{\lambda}, \quad (\lambda \text{ is the wavelength}). \tag{4}$$

$$\delta = \ln\left\{\frac{1}{4}\left[\sqrt{1+\left(\frac{a}{h-z}\right)^2}+1\right]\right.$$
$$\left.\cdot\left[\sqrt{1+\left(\frac{a}{h+z}\right)^2}+1\right]\right\}. \tag{5}$$

$$r_{11} = \sqrt{(z-z')^2 + a^2} \approx |z - z'|. \tag{6}$$

θ_{1k} ($k=2$, 3, 4, and 5) is the phase of the current in antenna k referred to the current in antenna (1).

$$r_{1k} = \sqrt{(z-z')^2 + b_{1k}^2}. \tag{7}$$

In (7), b_{1k} is the distance (taken in the meridian plane) between the center of antenna (1) and the center of antenna k.

Clearly, one may write

{Last four integrals in (1)}

$$= \sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} I_z' \frac{e^{-j\beta r_{1k}}}{r_{1k}}\, dz'. \tag{8}$$

Here n is the number of antennas in the array—five in the present case.

Thus, the current in antenna (1) of an n-element symmetrical array becomes

$$I_z = -\frac{j4\pi}{\Omega R_c}\left\{C_1 \cos \beta z + \tfrac{1}{2}V_o^e \sin \beta |z|\right\}$$
$$-\frac{1}{\Omega}\left\{I_z \ln\left(1-\frac{z^2}{h^2}\right) + I_z\delta\right.$$
$$\left.+\int_{-h}^{+h} \frac{I_z' e^{-j\beta r_{11}} - I_z}{r_{11}}\, dz'\right\}$$
$$-\frac{1}{\Omega}\sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} I_z' \frac{e^{-j\beta r_{1k}}}{r_{1k}}\, dz'. \qquad (9$$

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* Decimal classification: R125. Original manuscript received by the Institute, May 18, 1945.

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[1] Ronold King and Charles W. Harrison, Jr., "Mutual and self-impedance for coupled antennas," *Jour. Appl. Phys.*, vol. 15, pp. 481–495; June, 1944.

[2] Charles W. Harrison, Jr., "Mutual and self-impedance for collinear antennas," PROC. I.R.E., vol. 33, pp. 398–408; June, 1945. The following corrections should be made in this paper: The universal magnetic constant of space II, as defined by (7), should appear in lieu of π in the numerators of (20), (23), (63), (64), and (65). In the latter equation, II occurs twice.

In (13), $z \rightarrow l/2$. The denominator of the first bracketed term of (52) should read $\Omega R_c \cos(\beta l/2)$. Similarly, parenthesis should be used about each factor 1/2 appearing in (60b) and (62). The negative sign should be removed before jA_1^{II} in (73). Concerning the several tables that appear on page 404: The heading for the first one is $h=\lambda/4$; $\Omega=10$; $(h/a=75)$. For the second table, the heading is $h=\lambda/2$; $\Omega=10$; $(h/a=75)$.

In line 12 from the bottom of page 400, right-hand column, $z_1=0$.

[3] Charles W. Harrison, Jr., "On the calculation of the impedance properties of parasitic antennas array involving elements of finite radius," *Jour. Amer. Soc. Naval Eng.*, vol. 57, pp. 224–239; May, 1945, and p. 435; August, 1945.

[4] Charles W. Harrison, Jr., "On the distribution of current along asymmetrical antennas," *Jour. Appl. Phys.*, vol. 16, pp. 402–408; July, 1945.

[5] Charles W. Harrison, Jr., "A theory for three-element broadside arrays," to be published.

In solving this equation, a process of iteration is used. For instance, the expression[6]

$$(I_z)_o = -\frac{j4\pi}{\Omega R_c}\left\{C_1 \cos \beta z + \tfrac{1}{2} V_o^s \sin \beta \left| z \right|\right\} \qquad (10)$$

is conveniently selected as the zeroth-order approximation for the distribution of current along antenna (1), and this is substituted in the integrals when computing a first-order correction to the current. This procedure may be extended indefinitely, thus building up a series expansion for the current. Convergence of this series is assumed. In the present paper, only two terms in the expansion will be retained, as experience in analyzing numerically other antenna types proves this procedure to be satisfactory.

The current I_s, flowing at the input terminals of antenna (1) may be obtained from (9) by writing $z=0$ throughout. Thus,

$$I_o = -\frac{j4\pi}{\Omega R_c}\{C_1\} - \frac{1}{\Omega}\int_{-h}^{+h} \frac{I_z' e^{-j\beta l_{11}} - I_o}{l_{11}}\, dz'$$
$$-\frac{1}{\Omega}\sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} I_z' \frac{e^{-j\beta l_{1k}}}{l_{1k}}\, dz'. \qquad (11)$$

The current $I_h=0$ for $z=\pm h$. For this condition (9) becomes

$$0 = -\frac{j4\pi}{\Omega R_c}\left\{C_1 \cos \beta h + \tfrac{1}{2} V_o^s \sin \beta h\right\}$$
$$-\frac{1}{\Omega}\int_{-h}^{+h} I_z' \frac{e^{-j\beta d_{11}}}{d_{11}}\, dz'$$
$$-\frac{1}{\Omega}\sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} I_z' \frac{e^{-j\beta d_{1k}}}{d_{1k}}\, dz'. \qquad (12)$$

The notation

$$l_{11} = \sqrt{(z')^2 + a^2} \approx \left| z' \right| \qquad (13)$$
$$l_{1k} = \sqrt{(z')^2 + b_{1k}^2} \qquad (14)$$
$$d_{11} = \sqrt{(h - z')^2 + a^2} \approx \left| h - z' \right| \qquad (15)$$
$$d_{1k} = \sqrt{(h - z')^2 + b_{1k}^2} \qquad (16)$$

is used in (11) and (12).

Let the following shorthand be adopted:

$$F_1(o) \approx -\int_{-h}^{+h} \frac{\cos \beta u\, e^{-j\beta |u|} - 1}{|u|}\, du \qquad (17)$$

$$F_1(h) \approx -\int_{-h}^{+h} \cos \beta u \frac{e^{-j\beta |h-u|}}{|h - u|}\, du \qquad (18)$$

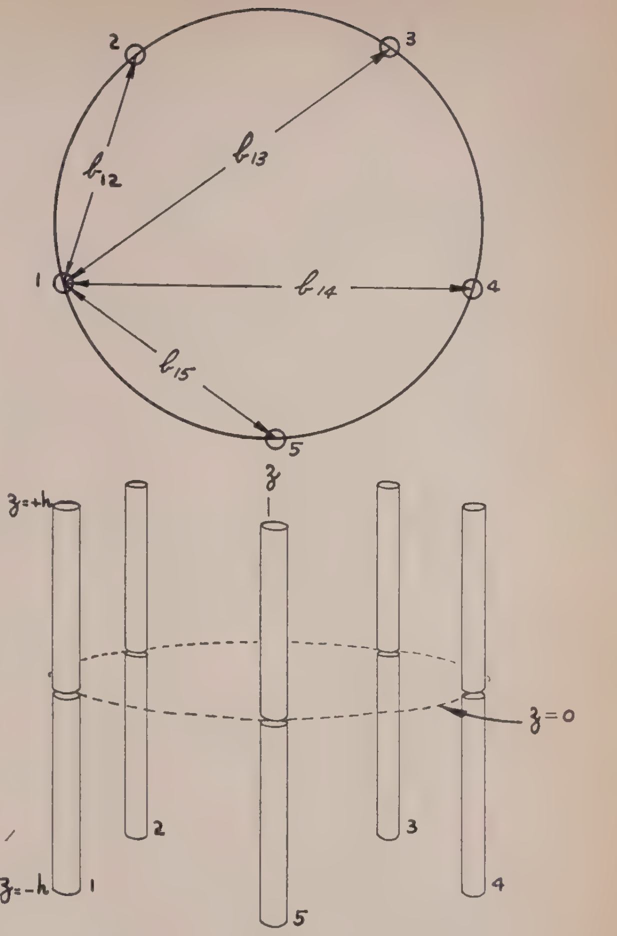

Fig. 1—Five-element circular array. All antennas have identical dimensions. They must be driven such that electrical symmetry obtains for the present theory to apply.

$$G_1(o) \approx -\int_{-h}^{+h} \sin \beta |u| \frac{e^{-j\beta |u|}}{|u|}\, du \qquad (19)$$

$$G_1(h) \approx -\int_{-h}^{+h} \sin \beta |u| \frac{e^{-j\beta |h-u|}}{|h - u|}\, du \qquad (20)$$

$$P_{1n}(o) = -\sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} \cos \beta u \frac{e^{-j\beta\sqrt{u^2+b_{1k}^2}}}{\sqrt{u^2+b_{1k}^2}}\, du \qquad (21)$$

$$P_{1n}(h) = -\sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} \cos \beta u \frac{e^{-j\beta\sqrt{(h-u)^2+b_{1k}^2}}}{\sqrt{(h-u)^2+b_{1k}^2}}\, du \qquad (22)$$

$$Q_{1n}(o) = -\sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} \sin \beta |u| \frac{e^{-j\beta\sqrt{u^2+b_{1k}^2}}}{\sqrt{u^2+b_{1k}^2}}\, du \qquad (23)$$

$$Q_{1n}(h) = -\sum_{k=2}^{k=n} e^{-j\theta_{1k}} \int_{-h}^{+h} \sin \beta |u| \frac{e^{-j\beta\sqrt{(h-u)^2+b_{1k}^2}}}{\sqrt{(h-u)^2+b_{1k}^2}}\, du. \qquad (24)$$

(Values for the above integrals are listed in Appendix I.)

Returning now to (11) and (12), and making use of (17) to (24), one obtains

$$I_o = -\frac{j4\pi}{\Omega R_c}\left\{C_1\left\{1 + \frac{1}{\Omega}\left(F_1(o) + P_{1n}(o)\right)\right\}\right.$$
$$\left. + \frac{1}{2} V_o^s \frac{1}{\Omega}\left\{G_1(o) + Q_{1n}(o)\right\}\right\} \qquad (25)$$

---

[6] The limitations of selecting (10) as the leading term in the antenna-current distribution are discussed in footnote reference 5. Readers should see these comments before applying the present theory to the solution of numerical problems.

The integral equation in the current for a symmetrical center-driven antenna is obtained from (9) by dropping the term involving a summation of integrals. If (10) is used in solving this equation, a limited investigation reveals that impedance values tend to be somewhat low for antennas operated in the vicinity of resonance, and high for antennas operated near antiresonance. The choice of (10) to represent the zeroth-order approximation for the current distribution, in lieu of the analytically superior form used in certain other papers, is admittedly an attempt to shorten the analysis at the expense of a some reduction in accuracy.

and

$$C_1 = -\frac{1}{2}V_o^e\left\{\frac{\sin\beta h + \frac{1}{\Omega}\left(G_1(h) + Q_{1n}(h)\right)}{\cos\beta h + \frac{1}{\Omega}\left(F_1(h) + P_{1n}(h)\right)}\right\}. \qquad (26)$$

Substituting the value of C_1 from (26) into (25), and neglecting terms in $1/\Omega^2$,

$$I_o = \frac{j2\pi V_o^e}{\Omega R_c}\left\{\frac{\sin\beta h + \frac{1}{\Omega}\left\{\sin\beta h(F_1(o) + P_{1n}(o)) - \cos\beta h(G_1(o) + Q_{1n}(o)) + G_1(h) + Q_{1n}(h)\right\}}{\cos\beta h + \frac{1}{\Omega}\left\{F_1(h) + P_{1n}(h)\right\}}\right\}. \qquad (27)$$

Since the input impedance Z_o of any antenna in the array is

$$Z_o = V_o^e/I_o \qquad (28)$$

$$Z_o = -j60\Omega\left\{\frac{\cos\beta h + \frac{1}{\Omega}\left\{F_1(h) + P_{1n}(h)\right\}}{\sin\beta h + \frac{1}{\Omega}\left\{\sin\beta h(F_1(o) + P_{1n}(o)) - \cos\beta h(G_1(o) + Q_{1n}(o)) + G_1(h) + Q_{1n}(h)\right\}}\right\}. \qquad (29)$$

Equation (29), together with values for $F_1(o)$, $F_1(h)$, etc., computed from (30) to (37), enables one to make a complete study of the impedance characteristics of an n-element symmetrical array. To obtain the impedance of base-driven vertical antennas of full length h located over a perfectly conducting plane earth, divide (29) by 2.

As a numerical illustration, assume that the array portrayed in Fig. 1 is to be analyzed. Let the following description apply: $h = \lambda/2$, $\Omega = 20$, $(h/a \approx 11{,}013)$, $b_{12}(=b_{15}) = \lambda/2$, $\theta_{1k} = 0$ degrees (all antennas are driven in phase). The object is to find the input impedance Z_o of antenna (1).

Since $n = 5$, $b_{13}\ \{=b_{14}\} = 0.8090\lambda$. From (31) to (37) one obtains

$$F_1(h) = -1.557 - j0.7461$$

$$G_1(o) = -1.418 + j2.438$$

$$G_1(h) = -0.6720 + j0.8810$$

$$P_{1n}(h) = -0.4868 - j0.8692$$

$$Q_{1n}(o) = 1.122 - j2.012$$

$$Q_{1n}(h) = -0.3310 - j1.073.$$

Substituting these values in (29)

$$Z_o = 4{,}998 - j19{,}465 \text{ ohms}.$$

This result, if collaborated by further evidence computed for different values of h and b_{12}, suggests that one property of circular arrays is high driving-point reactance, especially when electrically short radiators are used. If this is true, one would obviously not attempt to use such an array for long-wave transmission (where physical limitations preclude the use of electrically long antennas) because of the inherently high base voltage encountered for modest radiated powers.

If antennas numbered (2) to (5) are moved into the far zone with respect to antenna (1), so that coupling may be considered negligible, the input impedance of antenna (1), as computed from the present theory,[6] is

$$Z_{oo} = 5703 - j3320 \text{ ohms}.$$

The theory of symmetrical arrays is of use in analyzing certain multiple-wire transmission lines for radiation loss.[7] The present method of solution is more rigorous than methods based on "assumed currents."

Corner-Reflector Antennas[8,9]

A fundamental problem involved in the design of a corner-reflector antenna is the determination of the driving-point impedance of the primary radiator. Subject to certain approximations, this impedance may be calculated by the method of the previous section, provided the corner angle is π/n radians, where n is any positive integer.

Fig. 2 illustrates a corner-reflector antenna, while Fig. 3 shows the image orientation for several corner angles. The currents in the radiators denoted by + are oppositely directed from those represented by o.

For the theory of symmetrical arrays to apply rigorously, one must assume that the reflecting plates are perfectly conducting and infinite in extent. If the reflectors are constructed of copper or other metal of low resistivity, practically no power will be consumed in heat. Thus the assumption of zero surface impedance is an excellent approximation. Practical experience proves that reflectors having dimensions of several wavelengths on a side may be regarded as "infinite" when one is concerned with the calculation of the terminal impedance of the exciting antenna.

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[7] Charles W. Harrison, Jr., "On the pickup of balanced four-wire lines," Proc. I.R.E., vol. 30, pp. 517–518; November, 1942.

[8] J. D. Kraus, "The corner reflector antenna," Proc. I.R.E., vol. 28, pp. 513–519; November, 1940.

[9] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Son, New York, N. Y., 1944, p. 145.

As a numerical example of the theory of symmetrical arrays applied to corner-reflector antennas, consider the following system:

Corner angle = 90 degrees.

Primary radiator description: $h = \dfrac{\lambda}{4}$, $\Omega = 20$.

$D = 0.3535\lambda$. The object is to find Z_o.

Referring to Fig. (3), clearly

$$\begin{aligned} b_{12}\{= b_{14}\} &= \frac{\lambda}{2} \\ b_{13} &= 0.7071\lambda \\ \theta_{12} &= \pi \text{ radians.} \\ \theta_{13} &= 2\pi \text{ radians.} \\ \theta_{14} &= 3\pi \text{ radians.} \end{aligned}$$

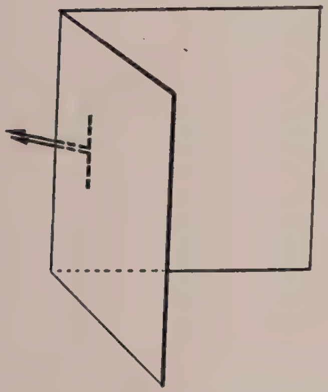


Fig. 2—A representative corner-reflector antenna.

For this case (30) to (37) give

$$\begin{aligned} F_1(o) &= 1.648 + j1.852 \\ F_1(h) &= -0.7090 + j1.219 \\ G_1(h) &= 1.815 + j1.143 \\ P_{1n}(o) &= -1.142 - j0.3457 \\ P_{1n}(h) &= -1.011 + j0.0073 \\ Q_{1n}(h) &= -0.9425 + j0.0326. \end{aligned}$$

Substituting these values in (29),

$$Z_o = 79.67 + j86.51 \text{ ohms.}$$

(If the reflecting plates are removed, the self-impedance of the antenna[6] is $Z_{oo} = 65.92 + j27.86$ ohms.)

The bandwidth of a corner-reflector antenna of fixed dimensions devolves into a study of (29) for appropriate increments in λ.

SYMMETRICAL ARRAY WITH CENTRAL ANTENNA

An interesting modification of the symmetrical array discussed earlier in the paper is achieved by the addition of a central radiator. Several such antenna systems are pictured in Fig. 4. The outer antennas, which lie equally spaced on the circumference of a circle, are identical. Normally they are required to carry currents of equal amplitude and phase. The dimensions of the central antenna need not correspond to those in the outer ring, and further, no restriction need be placed on the amplitude or phase of the current flowing therein.

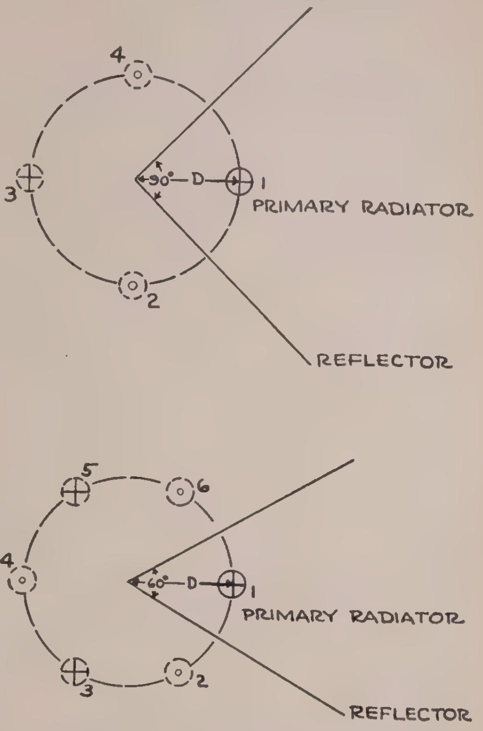
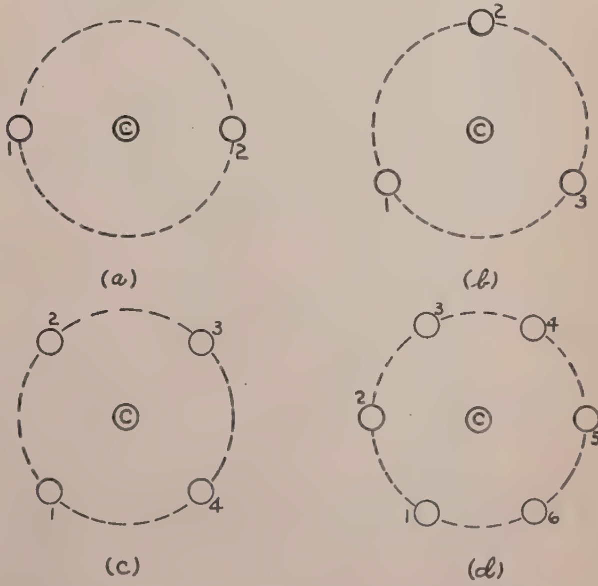


Fig. 3—Corner-reflector antenna showing image configuration for a 90-degree corner, and for a 60-degree corner.



(a) (b) (c) (d)

Fig. 4—Some symmetrical arrays with central radiators.

If the vector relationship between the applied voltages is known, one may determine the terminal impedance of any antenna in the array. Conversely, if the complex ratio of the driving-point currents is specified, the required geometrical relationship between the applied voltages is available. Suffice it to say, the solution of this problem is somewhat more involved than a quantitative discussion of symmetrical arrays, as two arbitrary constants, instead of one, must be determined.[5] One of these constants is associated with the distribution of current along the central radiator; the other is intrinsic in the expression for the current along any antenna in the outer ring.

Readers interested in analyzing a symmetrical array with central radiator may do so by synthesizing the first part of the present paper with the discussion previously presented.[5] In this paper, the array depicted in Fig. 4(a) is considered in detail. It is of interest to observe that, in lieu of an antenna at the geometrical center of a circular array, one may interpose another symmetrical array, and thus create two concentric rings of antennas. The increase in complexity of analysis is slight, provided symmetry obtains, both physical and electrical.

In conclusion, the writer wishes to mention an application of symmetrical-array theory to certain types of fixed antenna direction finders.[10] Such systems often consist of four antennas located at the corners of a square, with a fifth antenna erected at the geometrical center. One might be interested in calculating the input impedance of the central antenna, for equal load impedances connected across the terminals of the other antennas. This problem presents no great difficulties. The procedure is to regard the outer antennas as loaded parasites; i.e., a voltage $V_o^s = -I_o Z_L$ is applied to each, where Z_L is the complex load impedance. One then solves for the current at the center of the fifth antenna. The applied voltage, divided by this current, gives the required input impedance.

APPENDIX I

The integrals (17) to (24) have the following values:

$$F_1(o) \approx \overline{\text{Ci}}\, 2\beta h + j\, \text{Si}\, 2\beta h \tag{30}$$

$$\begin{aligned} F_1(h) \approx{}& \tfrac{1}{2} \cos \beta h \{ \overline{\text{Ci}}\, 4\beta h + j\, \text{Si}\, 4\beta h \} \\ &- \tfrac{1}{2} \sin \beta h \{ \text{Si}\, 4\beta h - j\, \overline{\text{Ci}}\, 4\beta h \} \end{aligned} \tag{31}$$

$$G_1(o) \approx -\,\text{Si}\, 2\beta h + j\, \overline{\text{Ci}}\, 2\beta h \tag{32}$$

$$\begin{aligned} G_1(h) \approx{}& -\tfrac{1}{2} \cos \beta h \{ \text{Si}\, 4\beta h - 2\, \text{Si}\, 2\beta h \\ &- j\, \overline{\text{Ci}}\, 4\beta h + j2\, \overline{\text{Ci}}\, 2\beta h \} \\ &- \tfrac{1}{2} \sin \beta h \{ \overline{\text{Ci}}\, 4\beta h + j\, \text{Si}\, 4\beta h - j2\, \text{Si}\, 2\beta h \\ &- 2\, \overline{\text{Ci}}\, 2\beta h - 4 \ln 2 \} \end{aligned} \tag{33}$$

$$\begin{aligned} P_{1n}(o) ={}& \sum_{k=2}^{k=n} e^{-j\theta_{1k}} \{ -\text{Ci}\, \beta(x_{1k} + h) + \text{Ci}\, \beta(x_{1k} - h) \\ &- j\, \text{Si}\, \beta(x_{1k} - h) + j\, \text{Si}\, \beta(x_{1k} + h) \} \end{aligned} \tag{34}$$

$$\begin{aligned} P_{1n}(h) ={}& \sum_{k=2}^{k=n} e^{-j\theta_{1k}} \{ -\tfrac{1}{2} \cos \beta h \{ \text{Ci}\, \beta(y_{1k} + 2h) \\ &- \text{Ci}\, \beta(y_{1k} - 2h) - j\, \text{Si}\, \beta(y_{1k} + 2h) \\ &+ j\, \text{Si}\, \beta(y_{1k} - 2h) \} + j\tfrac{1}{2} \sin \beta h \{ 2\, \text{Ci}\, \beta b_{1k} \\ &- j2\, \text{Si}\, \beta b_{1k} - \text{Ci}\, \beta(y_{1k} - 2h) - \text{Ci}\, \beta(y_{1k} + 2h) \\ &+ j\, \text{Si}\, \beta(y_{1k} - 2h) + j\, \text{Si}\, \beta(y_{1k} + 2h) \} \} \end{aligned} \tag{35}$$

$$\begin{aligned} Q_{1n}(o) ={}& \sum_{k=2}^{k=n} j e^{-j\theta_{1k}} \{ 2\, \text{Ci}\, \beta b_{1k} - j2\, \text{Si}\, \beta b_{1k} \\ &- \text{Ci}\, \beta(x_{1k} + h) - \text{Ci}\, \beta(x_{1k} - h) \\ &+ j\, \text{Si}\, \beta(x_{1k} + h) + j\, \text{Si}\, \beta(x_{1k} - h) \} \end{aligned} \tag{36}$$

$$\begin{aligned} Q_{1n}(h) ={}& \sum_{k=2}^{k=n} \frac{j}{2} e^{-j\theta_{1k}} [e^{-j\beta h} \{ 2\, \text{Ci}\, \beta(x_{1k} - h) \\ &- j2\, \text{Si}\, \beta(x_{1k} - h) - \text{Ci}\, \beta(y_{1k} - 2h) \\ &+ j\, \text{Si}\, \beta(y_{1k} - 2h) - \text{Ci}\, \beta b_{1k} + j\, \text{Si}\, \beta b_{1k} \} \\ &+ e^{j\beta h} \{ 2\, \text{Ci}\, \beta(x_{1k} + h) - j2\, \text{Si}\, \beta(x_{1k} + h) \\ &- \text{Ci}\, \beta(y_{1k} + 2h) + j\, \text{Si}\, \beta(y_{1k} + 2h) \\ &- \text{Ci}\, \beta b_{1k} + j\, \text{Si}\, \beta b_{1k} \}]. \end{aligned} \tag{37}$$

In the above

$$\text{Ci}\, z = \int_{\infty}^{z} \frac{\cos u}{u}\, du \tag{38}$$

$$\text{Si}\, z = \int_{0}^{z} \frac{\sin u}{u}\, du \tag{39}$$

$$\overline{\text{Ci}}\, z = \int_{0}^{z} \frac{1 - \cos u}{u}\, du = 0.5772 + \ln z - \text{Ci}\, z \tag{40}$$

$$x_{1k} = \sqrt{h^2 + b_{1k}^2} \tag{41}$$

$$y_{1k} = \sqrt{(2h)^2 + b_{1k}^2}. \tag{42}$$

ACKNOWLEDGMENT

Professor Ronold W. P. King, of Cruft Laboratory, Graduate School of Engineering, Harvard University, has reviewed the manuscript. The writer hopes to incorporate the material within a thesis at some future date.

[10] T. J. Keary, "Angles of Arrival of Radio Waves," Research Library of Physics, Harvard University, Cambridge, Mass., 1941.

General Formulas for "T"- and "Π"-Network Equivalents*

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***Summary*—This paper presents the development of two sets of general formulas which determine a set of "T" or "Π" impedances equivalent to any linear, lumped-constant, four-terminal network.**

IN SPITE of the many years that the equivalence of T and Π networks has been known and used, general formulas for the T and Π equivalents of four-terminal networks are not available. Such general formulas will be developed in this paper.

Consider, first, the general voltage equations of the four-terminal network[1] indicated in Fig. 1. In matrix form these equations are

$$\begin{vmatrix} V_1 \\ -V_2 \\ 0 \\ \cdot \\ \cdot \\ 0 \end{vmatrix} = \begin{vmatrix} Z_{11} & Z_{12} \cdots & Z_{1n} \\ Z_{21} & Z_{22} \cdots & Z_{2n} \\ Z_{31} & Z_{32} \cdots & Z_{3n} \\ \cdot & \cdot & \cdot \\ \cdot & \cdot & \cdot \\ Z_{n1} & Z_{n2} \cdots & Z_{nn} \end{vmatrix} \begin{vmatrix} I_1 \\ I_2 \\ I_3 \\ \cdot \\ \cdot \\ I_n \end{vmatrix}. \tag{1}$$

Inverting this equation leads to

$$\begin{vmatrix} I_1 \\ I_2 \\ I_3 \\ \cdot \\ \cdot \\ I_n \end{vmatrix} = \frac{1}{D} \begin{vmatrix} d_{11} & d_{12} \cdots & d_{1n} \\ d_{21} & d_{22} \cdots & d_{2n} \\ d_{31} & d_{32} \cdots & d_{3n} \\ \cdot & \cdot & \cdot \\ \cdot & \cdot & \cdot \\ d_{n1} & d_{n2} \cdots & d_{nn} \end{vmatrix} \begin{vmatrix} V_1 \\ -V_2 \\ 0 \\ \cdot \\ \cdot \\ 0 \end{vmatrix} \tag{2}$$

where D = the determinant of the impedance matrix
d_{ij} = the cofactor of the jth row and ith column of D.

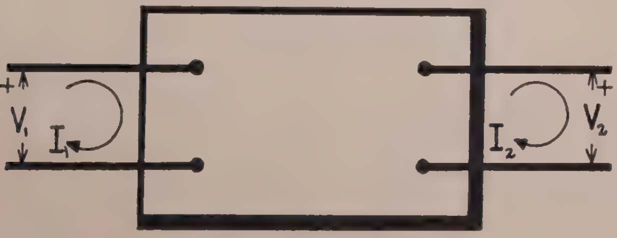

Fig. 1—General four-terminal network.

Because of the zeros appearing in the voltage matrix, the input and output currents of the four-terminal network are

$$\begin{vmatrix} I_1 \\ I_2 \end{vmatrix} = \frac{1}{D} \begin{vmatrix} d_{11} & d_{12} \\ d_{21} & d_{22} \end{vmatrix} \begin{vmatrix} V_1 \\ -V_2 \end{vmatrix}. \tag{3}$$

If, after multiplying the last of these current equations by -1, they are viewed as the node equations[2–4] of a three-node network (see Fig. 2) with the low node as the reference,

$$Z_1 = \frac{D}{d_{11} - d_{12}}$$
$$Z_2 = \frac{D}{d_{12}} = \frac{D}{d_{21}} \tag{4}$$
$$Z_3 = \frac{D}{d_{22} - d_{21}}$$

will make I_1 and I_2 of Figs. 1 and 2 the same for the same V_1 and V_2 provided $d_{12} = d_{21}$, as is actually the case. Consequently, the impedances of (4) specify a Π network equivalent to any linear, lumped-constant, four-terminal network.

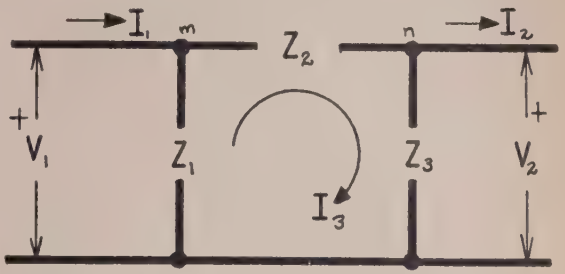

Fig. 2—General π network.

Formulas for a T equivalent to any linear, lumped-constant, four-terminal network can be established by inverting (3). Thus

$$\begin{vmatrix} V_1 \\ -V_2 \end{vmatrix} = \frac{D}{D'} \begin{vmatrix} d_{22} & -d_{21} \\ -d_{12} & d_{11} \end{vmatrix} \begin{vmatrix} I_1 \\ I_2 \end{vmatrix} \tag{5}$$

where

$$D' = \begin{vmatrix} d_{11} & d_{12} \\ d_{21} & d_{22} \end{vmatrix}. \tag{6}$$

But such a determinant consisting of cofactors of another determinant can be expressed as[5]

$$D' = \begin{vmatrix} d_{11} & d_{12} \\ d_{21} & d_{22} \end{vmatrix} = Dd_{22}'' \tag{7}$$

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† Bell Telephone Laboratories, New York, N. Y., on leave of absence from Illinois Institute of Technology.

[1] E. A. Guillemin, "Communication Networks," John Wiley and Sons, New York, N. Y., 1935, Vol. 2, Chapter IV.

[2] Myril B. Reed, "Node equations," Proc. I.R.E., vol. 32, pp. 355–359; June, 1944.

[3] Electrical Engineering Staff, Massachusetts Institute for Technology, "Electric Circuits," John Wiley and Sons, New York, N. Y. 1940, pp. 121–164, 391–398, 420–430.

[4] Murray F. Gardner and John L. Barnes, "Transients in Linear Systems," John Wiley and Sons, New York, N. Y., 1942, pp. 38 *et sequens*.

[5] M. Bocher, "Introduction to Higher Algebra," Macmillan Publishing Co., New York, N. Y., 1938, p. 33.

where d_{22}'' = the determinant left after removing the first two rows and first two columns of D.

Equation (5) thus becomes

$$\begin{vmatrix} V_1 \\ -V_2 \end{vmatrix} = \frac{1}{d_{22}''}\begin{vmatrix} d_{22} & -d_{21} \\ -d_{12} & d_{11} \end{vmatrix}\begin{vmatrix} I_1 \\ I_2 \end{vmatrix}. \qquad (8)$$

If this equation is considered as the two voltage equations of the T network of Fig. 3,

$$\begin{aligned} Z_a &= \frac{d_{11} - d_{12}}{d_{22}''} \\ Z_b &= \frac{d_{12}}{d_{22}''} = \frac{d_{21}}{d_{22}''} \\ Z_c &= \frac{d_{22} - d_{21}}{d_{22}''} \end{aligned} \qquad (9)$$

will make I_1 and I_2 of this T network the same as I_1 and I_2 of the four-terminal network from which the cofactors are determined if V_1 and V_2 are the same for both networks. Equations (9) accordingly determine the T network equivalent to any linear, lumped-constant, four-terminal network.

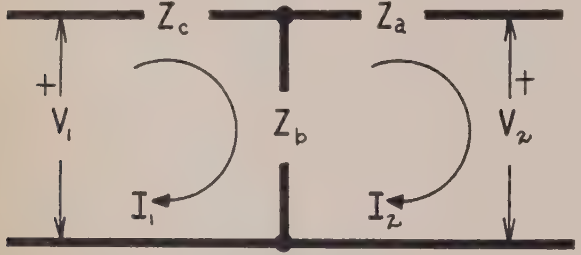

Fig. 3—General *T* network.

As an example of the use of the general formulas (4) and (9), the Π network equivalent to a T network can be determined from (4) as follows: The voltage equations for a T network (Fig. 3) are

$$\begin{vmatrix} V_1 \\ -V_2 \end{vmatrix} = \begin{vmatrix} Z_c + Z_b & -Z_b \\ -Z_b & Z_a + Z_b \end{vmatrix}\begin{vmatrix} I_1 \\ I_2 \end{vmatrix} \qquad (10)$$

and from these equations

$$\begin{aligned} d_{11} &= Z_a + Z_b \\ d_{12} &= d_{21} = Z_b \\ d_{22} &= Z_c + Z_b \\ D &= Z_aZ_b + Z_bZ_c + Z_cZ_a. \end{aligned} \qquad (11)$$

Substituting these last relations into (4) gives

$$\begin{aligned} Z_1 &= \frac{Z_aZ_b + Z_bZ_c + Z_cZ_a}{Z_a} \\ Z_2 &= \frac{Z_aZ_b + Z_bZ_c + Z_cZ_a}{Z_b} \\ Z_3 &= \frac{Z_aZ_b + Z_bZ_c + Z_cZ_a}{Z_c} \end{aligned} \qquad (12)$$

which are the well-known formulas for the Π impedances equivalent to a T network.

Similarly, for the Π network of Fig. 2, the voltage equations are

$$\begin{vmatrix} V_1 \\ -V_2 \\ 0 \end{vmatrix} = \begin{vmatrix} Z_1 & 0 & -Z_1 \\ 0 & Z_3 & -Z_3 \\ -Z_1 & -Z_3 & Z_1+Z_2+Z_3 \end{vmatrix}\begin{vmatrix} I_1 \\ I_2 \\ I_3 \end{vmatrix} \qquad (13)$$

from which, as required by (9)

$$\begin{aligned} d_{11} &= \begin{vmatrix} Z_3 & -Z_3 \\ -Z_3 & Z_1+Z_2+Z_3 \end{vmatrix} = Z_1Z_3 + Z_2Z_3 \\ d_{12} = d_{21} &= -\begin{vmatrix} 0 & -Z_1 \\ -Z_3 & Z_1+Z_2+Z_3 \end{vmatrix} = Z_1Z_3 \\ d_{22} &= \begin{vmatrix} Z_1 & -Z_1 \\ -Z_1 & Z_1+Z_2+Z_3 \end{vmatrix} = Z_1Z_2 + Z_1Z_3 \\ d_{22}'' &= Z_1+Z_2+Z_3. \end{aligned} \qquad (14)$$

Substituting these results into (9) gives

$$\begin{aligned} Z_a &= \frac{Z_2Z_3}{Z_1 + Z_2 + Z_3} \\ Z_b &= \frac{Z_1Z_3}{Z_1 + Z_2 + Z_3} \\ Z_c &= \frac{Z_1Z_2}{Z_1 + Z_2 + Z_3} \end{aligned} \qquad (15)$$

which are the well-known formulas for the T network equivalent to a Π network.

As a further example of the application of the general formulas (4) and (9), the Π and T equivalents of the network of Fig. 4 will be established. An examination of (4) and (9) shows that the determinant D, three of its cofactors d_{11}, $d_{12} = d_{21}$, and d_{22}, and the special subdeterminant d_{22}'' are required to determine these equations and so the equivalent Π and T networks.

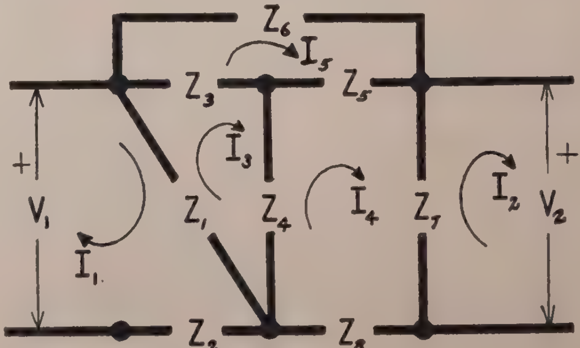

Fig. 4—Four-terminal network.

Writing the Kirchhoff equations for the network of Fig. 4 with the variables in the order of their subscripts, their coefficients are seen to give, for the determinant of the circuit,

$$D = \begin{vmatrix} Z_1+Z_2 & 0 & -Z_1 & 0 & 0 \\ 0 & Z_7 & 0 & -Z_7 & 0 \\ -Z_1 & 0 & Z_1+Z_3+Z_4 & -Z_4 & -Z_3 \\ 0 & -Z_7 & -Z_4 & Z_4+Z_5+Z_7+Z_8 & -Z_5 \\ 0 & 0 & -Z_3 & -Z_5 & Z_3+Z_5+Z_6 \end{vmatrix}. \tag{16}$$

From this determinant

$$d_{11} = \begin{vmatrix} Z_7 & 0 & -Z_7 & 0 \\ 0 & Z_1+Z_3+Z_4 & -Z_4 & -Z_3 \\ -Z_7 & -Z_4 & Z_4+Z_5+Z_7+Z_8 & -Z_5 \\ 0 & -Z_3 & -Z_5 & Z_3+Z_5+Z_6 \end{vmatrix} \tag{17}$$

$$d_{12} = d_{21} = -\begin{vmatrix} 0 & -Z_1 & 0 & 0 \\ 0 & Z_1+Z_3+Z_4 & -Z_4 & -Z_3 \\ -Z_7 & -Z_4 & Z_4+Z_5+Z_7+Z_8 & -Z_5 \\ 0 & -Z_3 & -Z_5 & Z_3+Z_5+Z_6 \end{vmatrix} \tag{18}$$

$$d_{22} = \begin{vmatrix} Z_1+Z_2 & -Z_1 & 0 & 0 \\ -Z_1 & Z_1+Z_3+Z_4 & -Z_4 & -Z_3 \\ 0 & -Z_4 & Z_4+Z_5+Z_7+Z_8 & -Z_5 \\ 0 & -Z_3 & -Z_5 & Z_3+Z_5+Z_6 \end{vmatrix} \tag{19}$$

$$d_{22}'' = \begin{vmatrix} Z_1+Z_3+Z_4 & -Z_4 & -Z_3 \\ -Z_4 & Z_4+Z_5+Z_7+Z_8 & -Z_5 \\ -Z_3 & -Z_5 & Z_3+Z_5+Z_6 \end{vmatrix}. \tag{20}$$

If these determinants are combined in accordance with equations (4) and (9), the impedance elements of an equivalent Π and equivalent T network will result. In particular, if the impedances of Fig. 4 are $Z_1=2+j2$, $Z_2=3+j4$, $Z_3=3-j4$, $Z_4=5-j5$, $Z_5=8+j6$, $Z_6=6-j8$, $Z_7=1-j1$, $Z_8=2-j4$, the determinant is

$$D = \begin{vmatrix} 5+j6 & 0 & -2-j2 & 0 & 0 \\ 0 & 1-j1 & 0 & -1+j1 & 0 \\ -2-j2 & 0 & 10-j7 & -5+j5 & -3+j4 \\ 0 & -1+j1 & -5+j5 & 16-j4 & -8-j6 \\ 0 & 0 & -3+j4 & -8-j6 & 17-j6 \end{vmatrix}$$

$D=4704-j21{,}472.$

Similarly,

$$d_{11} = -1456 - j2592$$

$$d_{12} = d_{21} = -412 + j516$$

$$d_{22} = 14{,}544 - j10{,}042$$

$$d_{22}'' = 548 - j2314.$$

The Π equivalent of Fig. 4 can now be determined by substituting these last results into (4); i.e.,

$$Z_1 = \frac{D}{d_{11}-d_{12}} = \frac{4704-j21{,}472}{-1044-j3108} = 5.75 + j3.45$$

$$Z_2 = \frac{D}{d_{12}} = \frac{4704-j21{,}472}{-412+j516} = -29.9 + j14.7$$

$$Z_3 = \frac{D}{d_{22}-d_{21}} = \frac{4704-j21{,}472}{14{,}956-j10{,}558} = 0.884 - j0.809.$$

Placing these impedances in the positions indicated in Fig. 2 establishes the Π equivalent of Fig. 4. Note that Z_2 is not physically realizable; consequently, for the numerical impedances specified in the foregoing, Fig. 4 does not have a physically realizable Π equivalent.

The T equivalent of Fig. 4 follows from evaluating equations (9) in terms of the numerical values established for d_{11}, d_{12}, d_{22}, and d_{22}''. Thus,

$$Z_a = \frac{d_{11}-d_{12}}{d_{22}''} = \frac{-1044-j3108}{548-j2314} = 1.17 - j0.729$$

$$Z_b = \frac{d_{12}}{d_{22}''} = \frac{-412+j516}{548-j2314} = -0.251 - j0.118$$

$$Z_c = \frac{d_{22}-d_{21}}{d_{22}''} = \frac{14{,}956-j10{,}558}{548-j2314} = 5.74 + j5.09.$$

Therefore, since R_b is negative, the four-terminal network of Fig. 4, for the numerical values of impedances specified in the foregoing, does not have a physically realizable T equivalent, either. This is often true also, when an equivalent T or Π is calculated from open-circuited and short-circuited impedance measurements on an actual circuit. Even physically realizable T or Π networks do not always have physically realizable Π or T equivalents.

Institute News and Radio Notes

Plans Go Forward for Winter Technical Meeting and Radio Engineering Show

PROGRESS on plans for the first postwar Winter Technical Meeting and Radio Engineering Show of the Institute at the Hotel Astor, New York, January 23 to 26, 1946, are far advanced, and all indications point to one of the largest as well as one of the most significant gatherings of this type ever held.

Last year, more than 3000 members were present and reports this year indicate a substantially greater attendance. Several features in addition to the major highlights of the Meeting are on the schedule of events.

In preparation for the Radio Engineering Show, 132 exhibitors have already taken the total of 168 booths originally planned, including three theater booths, while efforts are being made to obtain additional space to accommodate the waiting list of further exhibitors. This mammoth show is scheduled to open at 4:00 P.M. Wednesday, January 23, and will close promptly at 2:00 P.M. Saturday, January 26.

The annual banquet, on Thursday, January 24, the social highlight of the I.R.E. year, will have places for 2500 members. Mr. Edgar Kobak, president of the Mutual Broadcasting System, will be the toastmaster. The principal speaker at the banquet will be Dr. Frank B. Jewett, president of the National Academy of Sciences.

For the President's Luncheon to be held Friday, January 25, it has been announced that Mr. L. M. Clement, vice-president in charge of research and engineering of the Crosley Corporation, will be master of ceremonies.

Another enjoyable feature this year will be a cocktail party also scheduled for Friday evening, for members and their friends. A nominal fee will be charged for admission.

An interesting program has been arranged for the women guests at the meeting. This will consist of sightseeing trips to points of interest in New York City and luncheons at unusual places are planned.

This year the I.R.E. will be host at a joint meeting with the American Institute of Electrical Engineers, scheduled to be held in the Engineering Society's auditorium on Wednesday evening, January 23. Last year, this meeting drew such crowds that many had to be turned away. At this gathering, however, arrangements have been made to install a public-address system and to reserve another large meeting room in the same building to accommodate any overflow attendance. At the joint meeting, there will also be a timely address by a speaker prominent in the electrical and electronics field.

Since the demand for hotel accommodations in New York is still extremely critical, those who have not as yet sent in their reservations for hotel rooms should do so at once.

Board of Directors

October 3 Meeting: At the October 3, 1945, meeting of the Board of Directors, the following were present: W. L. Everitt, president; G. W. Bailey, executive secretary; S. L. Bailey, E. F. Carter, Alfred N. Goldsmith, editor; R. F. Guy, R. A. Hackbusch, R. A. Heising, treasurer; Keith Henney, F. B. Llewellyn, Haraden Pratt, secretary; D. B. Sinclair, W. O. Swinyard, H. M. Turner, H. A. Wheeler, and W. C. White.

Approval of Executive Committee Actions: It was unanimously approved that the actions of the Executive Committee taken at its September 5, 1945, meeting be ratified.

Constitution and Laws Committee: The Board approved the adoption of the following proposed modifications of Article III, Section 2, and Article II, Section 6:

> "ARTICLE III, SECTION 2—Applications for admission or transfer to any grade of membership, except Fellow, shall be addressed to the Board of Directors and submitted to the Institute office. Election or transfer of an applicant to any grade, except Fellow, shall be made by a two-thirds affirmative vote of the Board membership voting, or a two-thirds affirmative vote of a committee of Board members, duly authorized in the By-laws to act for the Board in electing members or transferring their membership grades."
>
> "ARTICLE II, SECTION 6—For admission to the grade of Student, a candidate shall be devoting a major portion of his time as a registered student in engineering or science in a school of recognized standing. Membership in this grade shall not extend more than two and one-half years beyond the termination of his student status as described above. A Student member who for any reason has not transferred to Member grade within the extended period shall be transferred to Associate grade."

ARTICLE IV, SECTION 2: Mr. Guy, chairman of the Constitution and Laws Committee, moved that the Minutes of the October 3, 1945, Board meeting show that the Board reconsidered Article IV, Section 2, and reiterated its decision as shown at the September 5, 1945, meeting. This was approved.

Section Redistricting: It was unanimously approved that the plan of the Section redistricting be accepted as submitted, with the recommendation that, when the map is published, United States territories only be shown (with a notation to the effect that Canadian territories will be shown at a later date).

During the discussion on the Section redistricting, Mr. R. A. Heising, chairman of the Sections Committee, was instructed to reopen, with the Connecticut Valley and Boston Sections, the discussion on the boundaries.

The change in the Sections territories is to become effective as of January 1, 1946.

Joint Committees: A motion that the Standards Committee and the Technical Committees be instructed to form where indicated, as speedily as possible, joint committees with other responsible technical organizations or groups dealing within their scope, and to carry forward the activities of such joint committees, was unanimously approved.

Bylaw Section 57: The price of a single annual subscription to the PROCEEDINGS was increased to $6.00 from $5.00, and the Constitution and Laws Committee was instructed to modify Bylaw Section 57 in accordance with this increased price.

Annual Meeting: The annual meeting of the Institute will be held on Thursday, January 24, 1946.

Radio Technical Planning Board: Mr. Pratt, Chairman of the RTPB, reported that the Federal Communications Commission was in favor of the continuation of the RTPB functions. Mr. Pratt also stated that it was opinion The Institute should take a leading part in the future activities of the RTPB.

Executive Committee

October 3 Meeting: At the Executive Committee meeting, held on October 3, 1945, the following were present: W. L. Everitt, president; G. W. Bailey, executive secretary; S. L. Bailey, W. H. Crew, assistant secretary; Alfred N. Goldsmith, editor; R. A. Heising, treasurer; and Haraden Pratt, secretary.

Membership Approval was given to the 323 applications for membership in the Institute listed on page 36A of the November, 1945, issue of the PROCEEDINGS. These applications are as follows:

For Transfer to Senior Member grade	36
For Admission to Senior Member grade	9
For Transfer to Member grade	62
For Admission to Member grade	63
For Admission to Associate grade	109
For Admission to Student grade	44
Total	323

Section Redistricting: It was unanimously approved that the Executive Committee recommend to the Board that the redistricting of the Sections, which has been done by the Sections Committee in compliance with the Board's instructions of March 7, 1945, be accepted.

At the suggestion of Editor Goldsmith, the map showing the redistricting of the Sections will be published in the PROCEEDINGS, together with appropriate explanations. Coincident with the publication of the map, an article covering the procedure by the Sections in handling higher grade membership transfers and admissions will be published.

Winter Technical Meeting: Definite arrangements have been completed with the Astor Hotel for the Winter Technical Meeting, and the budget, as set up, was unanimously approved.

Postwar Publication Fund: Dr. Goldsmith moved that Dr. F. E. Terman, Chairman, Dr. B. E. Shackelford, and Mr. I. S. Coggeshall be appointed a committee, with whom Dr. Goldsmith will collaborate to secure an addition to the Postwar Publication Fund for further expansion of technical publications.

Forthcoming Section Meetings

December 13, 1945

CONNECTICUT VALLEY
BRIDGEPORT

Dinner: 6:45 P.M., Stratfield Hotel
Meeting: 8:00 P.M., Stratfield Hotel
Chairman: H. W. Sundius
Speaker: A. E. Harrison, Sperry Gyroscope Company
Title of Paper: The Klystron

December 18, 1945

CINCINNATI

Meeting: Engineering Society of Cincinnati Headquarters
Speaker: H. B. Fancher, General Electric Company
Title of Paper: Microwave Relay Networks

December 21, 1945

CHICAGO

Dinner: 6:00 P.M., Main Dining Room, 38th Floor, Civic Opera Building, 20 Wacker Drive, Chicago, Illinois
Program: Sixth Floor, Civic Opera Building
Chairman: Ralph P. Glover
Speaker: Hugh S. Knowles, Vice-President and Chief Engineer, Jensen Radio Manufacturing Co.
Title of Paper: Receiver Loudspeaker Systems

Summary—Historical data on radio receiver loudspeaker systems will be given which indicate certain prewar trends. Stress is placed on the deliberate use of distortion of various kinds in narrow-band transmission systems and the limitations which similar distortion place on wide-band systems. The limitations of single-channel, wide-band transmission systems will be discussed in relation to the limitations to be expected from the acoustic portions of frequency-modulation systems. War developments in loudspeakers will be included for what light they may shed on their influence on postwar design trends. A demonstration will be included.

January 8, 1946

CONNECTICUT VALLEY
SPRINGFIELD

Dinner: 6:45 P.M., Hotel Sheraton
Meeting: 8:00 P.M., Hotel Sheraton
Chairman: H. W. Sundius
Speaker: H. B. Fancher, General Electric Company
Title of Paper: General Electric-International Business Machines Radio Relay

January 14, 1946

PITTSBURGH

Meeting: 8:00 P.M., Mellon Institute
Students from the University of Pittsburgh and Carnegie Institute of Technology will present papers.

January 15, 1945

CINCINNATI

Meeting: Engineering Society of Cincinnati Headquarters
Speaker: John D. Reid, Crosley Corporation
Title of Paper: Sound and Hearing

January 17, 1946

CONNECTICUT VALLEY
HARTFORD

Dinner: 6:45 P.M., Hotel Bond
Meeting: 8:00 P.M., Hotel Bond
Chairman: H. W. Sundius
Speaker: Daniel E. Noble, Galvin Manufacturing Company
Title of Paper: The Motorola Walkie Talkie, Handie Talkie, and the New 152- to 160-Megacycle Portable Mobile Equipment

* * *

NOTE: We are desirous of giving as much publicity as possible to forthcoming Section Meetings, provided that we have the material in sufficient time to include it in the next available issue of the PROCEEDINGS. For the February, 1946, issue, we should have copy not later than December 20, for March, not later than January 20, and so on.

The Editor

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Election of Officers

On November 13, 1945, the Board of Directors announced that Dr. Frederick B. Llewellyn was elected president of The Institute of Radio Engineers for 1945, and Mr. E. M. Deloraine was elected vice-president. The following directors also were chosen by the membership: Dr. Walter G. R. Baker, Mr. Virgil M. Graham, and Dr. David B. Sinclair.

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Kansas City Technical Societies Council

Seventeen technical and professional societies of Greater Kansas City recently formed the Technical Societies Council of Kansas City. This organization is to be a co-ordinating agency which will further the industrial upbuilding of the region. The Institute of Radio Engineers is represented by R. N. White (S'40-A'41-M'45), engineer and maintenance department, T.W.A., and A. P. Stuhrman (A'39), of the Wilcox Electric Company.

I.R.E. People

J. B. Coleman

John B. Coleman and M. C. Batsel

Appointment of John B. Coleman (A'25-M'29-SM'43) as assistant director of engineering for the RCA Victor Division and the naming of M. C. Batsel (A'21-F'27) as chief engineer of the engineering products have been announced by D. F. Schmit (A'25-M'38-SM'43) director of engineering for the RCA Victor Division.

Mr. Coleman, who will make his headquarters at the company's home office in Camden, N. J., joined RCA in 1930. He progressed through the engineering division holding a series of responsible positions until 1939 when he was appointed chief engineer of the engineering products department, a position he held until his new assignment.

Mr. Batsel is widely known in the radio and motion-picture industry. He became associated with RCA in 1929. Previous to his new assignment, Mr. Batsel was chief engineer at the RCA Victor plant in Indianapolis, Ind.

M. C. Batsel

◆

Donald G. Fink

Donald G. Fink (A'35-SM'45) spoke before the Radio Club of America on September 13, 1945, on "Radar Theory and Wartime Applications." This paper dealt with the fundamental basis of radar detection as derived from simple concepts and expressed in the form of a single equation which expresses the effects of transmitter power and receiver noise, target size and distance from the transmitter, wavelength, and antenna characteristics. The application of the equation to the design of radar equipment was discussed and illustrated with slides of typical military and naval radars, together with the technical background of wartime applications.

Browder J. Thompson Memorial

In the May Proceedings, an announcement was made of a plan to turn contributions of friends of the late B. J. Thompson over to The Institute of Radio Engineers to establish an annual prize for the outstanding paper in radio and electronics.

Undoubtedly many members of the Institute will be interested in the status of the Browder J. Thompson Memorial Fund. As of October third, contributions totaling over $4100 have been received.

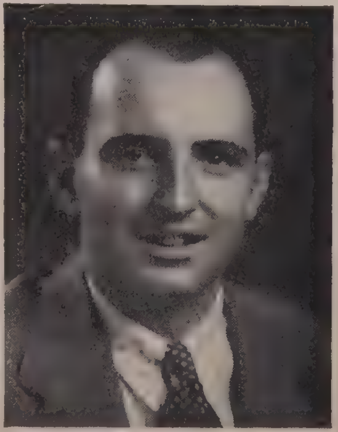
The Fund is still growing; and inasmuch as only about one fourth of the potential contributors have been heard from, there is every reason to believe that it will continue to grow.

In view of this favorable preliminary result, consideration is being given to investing the money now on hand and taking other appropriate steps in order that the initial award may be made at an early date, possibly at the forthcoming Summer Convention. Naturally, B. J.'s friends are strongly desirous that the fund established in his name shall speedily encourage workers in the field with which he was so closely identified and to which he contributed so greatly.

R. R. Law
Secretary
Memorial Committee

Mr. Fink is a graduate of the Massachusetts Institute of Technology, and spent a year in graduate study before joining the editorial staff of *Electronics* in 1934. After a four-year leave of absence, Mr. Fink returned to *Electronics*, on September 1, 1945, as executive editor. During his leave of absence, he was associated with the Radiation Laboratory, assigned to development work on the Loran system of long-range navigation. The Loran transmitters now in use are based on his designs, and he is the author of "Microwave Radar," the first radar textbook, written for the use of the laboratory staff and military personnel.

Appointed head of the Loran division of the laboratory in 1943, Mr. Fink served both in European and Pacific areas, and was later transferred to the Washington, D. C., office of the Secretary of War, where he advised the Army Air Forces concerning the use of Loran and related navigational devices. Assigned to the headquarters of the Ground Forces, Mr. Fink advised on the use of radar as applied to gunfire control and the detection of ground targets. He was also appointed a member of the Committee on Air Navigation and Traffic Control, and will continue this government service in conjunction with his new position.



Donald G. Fink

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RMA–I.R.E. Golf Tournament

The Radio Manufacturers Association of Canada and the Toronto Section of The Institute of Radio Engineers held their first joint golf tournament on September 12, 1945. The I.R.E. trophy for low gross was won by E. O. Swan (A'40–M'44).

The tournament was followed by a dinner, at which J. R. Longstaffe (A'36), chairman of the golf committee, presided. Mr. Longstaffe introduced several representatives of both organizations, including F. H. R. Pounsett (A'26–SM'44), chairman of the Toronto Section, and Alexander Bow (A'43), secretary-treasurer of the Toronto Section. Mr. Pounsett spoke briefly on the great strides which the art of radio had made during the war years, and predicted that radio will be *the* industry in the future.

Books

The Decibel Notation and Its Applications to Radio Engineering and Acoustics, by V. V. L. Rao

Published (1944) by Addison and Company, Inc., Madras, India. 176 pages+3-page index + xvi pages. 51 illustrations. 8½×5¾ inches. Price, $3.00.

The decibel notation may be taken to mean the use in engineering of the logarithm of the ratio of two powers, or of their associated quantities, rather than the ratio itself. As such it must pervade any book on communication engineering. The present work however makes it the focal point.

The first part presents the reasons for using logarithms, and develops the concepts underlying those quantities such as loss, power level, etc., which occur in electrical systems and are expressed in decibels. The second part deals with acoustic systems, and includes the concept of the phon, which replaces the decibel when dealing with loudness level. The third and largest part is devoted to specific numerical examples of the calculation, in decibels, of the various quantities which are met in the engineering of radio and acoustic systems. Methods of measuring these are also described. Appendixes include tables and a discussion of the use of logarithmic graphs.

The level of difficulty has been chosen to fit "an average student of electrical engineering." The practicing engineer would find it useful primarily as a reference. It should also meet the needs of those in administrative positions whose duties require a working knowledge of the language of the engineer.

The style is clear and straightforward, and as Professor Chakravarti says in his foreword, "This modest work is packed with much useful and worthwhile information presented with clarity and economy of words."

Reporting as it does the current terminology of communication it illustrates how far the application of the decibel has evolved from its original use as a measure of "transmission loss" which described the change in received power in going from the reference to the test conditions. Also it reflects that looseness in the use of terms which is so prevalent in the literature.

R. V. L. Hartley
Bell Telephone Laboratories
Murray Hill, N. J.

On the Propagation of Radio Waves, by Olof E. H. Rydbeck

Published (1944) by Elanders Boktryckeri Aktiebolag, Göteborg, Sweden, 169 pages. 47 illustrations. 7×10 inches. Price, 10.0 kronor.

In this book the author obtains theoretical solutions to several problems connected with the idealized problem of the propagation of radio waves between the earth and a concentric parabolic layer of ionization. The book is addressed primarily to the mathematical physicist and unquestionably constitutes a contribution in this field by presenting explicit solutions of many interesting problems encountered in studying the transmission of radio waves in inhomogeneous media. The exact solution of the problem of determining the intensities of radio waves at large distances from the transmitting antenna in the presence of an ideally smooth earth and upper atmosphere should provide a useful guide to the calculation of the expected intensities of ionospheric waves as actually propagated around the earth. Previous attempts to solve this problem have almost always used the methods of geometrical optics and this paper indicates in what frequency and distance ranges such methods may be expected to be applicable.

In addition formulas are given for the expected sky-wave (and ground-wave) field intensities to be expected (in the presence of the idealized earth and atmosphere) in many cases where the methods of geometrical optics would not be applicable. In one case, i.e., transmission on a frequency of 60 kilocycles over sea water, these solutions are actually reduced to the form of a curve showing the attenuation factor as a function of distance. This curve indicates that the field intensities to be expected oscillate up and down with increasing distance in a very irregular way. Unfortunately no experimental data were given to confirm this rather unexpected behavior and it is the belief of the reviewer that the actual transmission on this frequency is not characterized by any such oscillations. Probably much additional theoretical work of the type contained in this book will be required before suitable theoretical solutions of this wave propagation problem are obtained.

Kenneth A. Norton
Office of the Chief Signal Officer
Washington 25, D. C.

Radio Pioneers' Books

Commemorating the Radio Pioneers' Dinner held in New York City on November 8, 1945, at the Hotel Commodore, the New York Section of The Institute of Radio Engineers, prepared a booklet containing historical notes on the I.R.E., Veteran Wireless Operators' Association, Radio Club of America, and the American Radio Relay League. This volume contains much of interest regarding radio from the mid 1800's through 1925, both chronologically and by eras. In it there are biographical sketches of numerous important personages in radio and a substantial number of photographs of interesting old equipment. The book was prepared by an editorial group under the chairmanship of Harold P. Westman. It contains 64 pages, is 8½ by 11 inches in size, attractively bound in a light-gray cover with blue printing, and is priced at $1.00 per copy. Copies of this book may be ordered from the chairman of the New York Section, Professor G. B. Hoadley, 85 Livingston St., Brooklyn 2, New York. Remittances should accompany orders.

UHF Radio Simplified, by Milton S. Kiver

Published (1945) by D. Van Nostrand Co., Inc., 250 Fourth Ave., New York, N. Y. 236 pages+2-page index+viii pages. 157 illustrations. 5¾×8½ inches. Price, $3.25.

In this book the author attempts a simplified, nonmathematical treatment of ultra-high-frequency radio, intended apparently for the use of radio amateurs; laboratory and technical assistants; technical, sales, and commercial representatives; executives; and others who have some technical knowledge and experience, who would profit by a simple descriptive treatment of the subject. The author assumes that the reader has some knowledge and experience with lower-frequency radio circuits, as, for example, broadcast transmitters and receivers, and attempts to explain the generation and transmission of ultra-high-frequency radiations, by pointing out similarities and analogies and indicating logical extensions of the lower-frequency techniques to the ultra-high-frequency region. Attempts are made to visualize the mechanism of electron flow and in other ways to develop physical concepts of phenomena in ultra-high-frequency generators and transmission lines.

On the whole, the author has been successful in accomplishing his stated purpose, namely, "to point out the underlying equality of all radio." The book undoubtedly will be of interest and value to the readers for which it is designed. It would be expected that the avoidance of mathematical analysis and the attempts to make the explanation easily comprehended would result in glaring omissions and a sense of incompleteness. These would, however, be felt more keenly by serious and thorough students in the field, for whom the book is not intended and to whom it is not recommended. These simplifications and the brevity of the text likewise cause the author to include categorical statements without indicating the basis for them or deriving them mathematically.

Following an introductory chapter, in which the transition from the lower- to the higher-frequency techniques is explained, the book describes the operation and use of generators of high-frequency oscillations; the transmission of high-frequency currents from the generator to the antenna; antennas; the propagation of high-frequency electromagnetic waves; and the instruments and methods of measurement of wavelengths, potential difference, and power. There is no adequate treatment of transmitter stations nor of receivers.

Advances in this field during the war have been rapid on account of the development of radar and special communication systems for military purposes. This new knowledge, as yet not released from military security, could not be included in the book. However, the ideas developed and the techniques described will continue to be important as an introduction to the more advanced ideas and techniques that will become available when military security permits.

O. S. Duffendack
North American Philips Company, Inc.
Irvington, N. Y.

Correspondence

Correspondence on both technical and nontechnical subjects from readers of the PROCEEDINGS OF THE I.R.E. is invited subject to the following conditions: All rights are reserved by the Institute. Statements in letters are expressly understood to be the individual opinion of the writer, and endorsement or recognition by the I.R.E. is not implied by publication. All letters are to be submitted as typewritten, double-spaced, original copies. Any illustrations are to be submitted as inked drawings. Captions are to be supplied for all illustrations.

Change of Member Address for 1946 Yearboook

If you have made any *changes* in your *address* or *position* since you sent in your YEARBOOK postcard questionnaire, will you please inform I.R.E. Headquarters promptly. It would be helpful in that case if you would make a statement to the effect that your YEARBOOK listing should be changed as follows: (Here insert the proper changes.) No YEARBOOK corrections in address or position can be made after January 1, 1946. Kindly address your communications to

The Institute of Radio Engineers, Inc.
330 West 42nd Street
New York 18, New York

Iron Powder Cores and Coils

When working with iron-core direction-finding loop antennas some time ago, the writer was greatly assisted in his understanding of the related permeability phenomena by the paper by R. M. Bozorth and D. M. Chapin.[1]

A perusal of articles which have since been published in various journals does not show any reference to this paper but many repeat earlier "rule-of-thumb" methods. To this writer the paper seems worthy of some elaboration in respect to its application to the design of iron powder cores and coils for radio sets.

Formula (1) for the effective permeability of a core and coil combination is derived from fundamental principles

$$\mu_e = \frac{L_i}{L_a} = \frac{d^2}{D^2}(\mu - 1) + 1 \qquad (1)$$

$$\frac{1}{\mu} = \frac{1}{\mu'} + \frac{N}{4\pi} \qquad (2)$$

and is accurate for an infinite coil and core length.[2]

SYMBOLS

μ_e = effective permeability of coil and core combination
μ = apparent peremability of cylindrical core
μ' = ring permeability of core material[3]
$N/4\pi$ = demagnetizing factor
L_i = inductance of coil with core
L_a = inductance of coil less core
$D = \dfrac{D_1 + D_2}{2}$
D_1 = outer diameter of the winding
D_2 = inner diameter of the winding
d = diameter of core.
l = length of core.

What to do when finite lengths and small length-to-diameter ratios are encountered is the problem constantly facing the designer. The usual method is to allow certain empirical factors known from experience or from scant published references to obtain the apparent permeability of various geometrical core configurations.

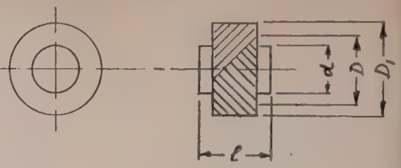


Fig. 1

The abovementioned article appears to clear this matter up for the so called "open construction" shown in Fig. 1. Using (2) which is rewritten from reference 1 and picking values from the curves for cylindrical rods in Fig. 2 on page 322 the following table is obtained.

m	ring permeability			
	5	10	100	infinity
1	2.25	2.90	3.2	3.5
2	3.29	4.65	6.5	7.6
3	3.85	5.92	10.08	12.3
4	4.22	6.94	15.0	17.7
5	4.46	7.69	20.0	24.0
6	4.59	8.15	25.0	31.0
8	4.74	8.75	34.0	46.0
10	4.83	9.14	41.6	62.5

The apparent permeability is found at the intersection of the column for the ring permeability of the material and the row for the appropriate core length-to-diameter ratio.

This table confirms some of the empirical figures used for the popular short cores with an *m* in the neighborhood[4] of 2. It is also quite evident that it is not economical to use high-permeability materials when *m* is small.

Thus this table, when used in conjunction with (1) rounds out the picture for the designer.

Conversely, it is possible to make measurements of the material permeability using ordinary coils and cores.

H. W. JADERHOLM
Montreal, Que., Canada

---

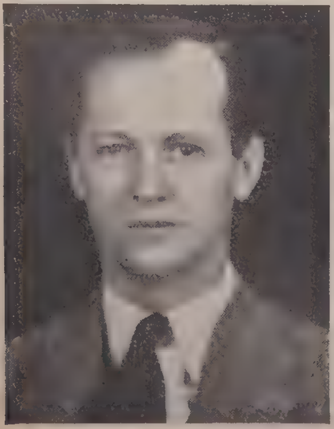
[1] R. M. Bozorth and D. M. Chapin, "Demagnetizing factors of rods," *Jour. Appl. Phys.*, p. 320; May, 1942.

[2] Dudley E. Foster and Arthur E. Newlon, "Measurements of iron cores at radio frequencies," PROC. I.R.E., vol. 29, pp. 266–277; May, 1941.

[3] At the same density.

[4] *m* = ratio of core length to diameter.

Contributors



W. S. BACHMAN

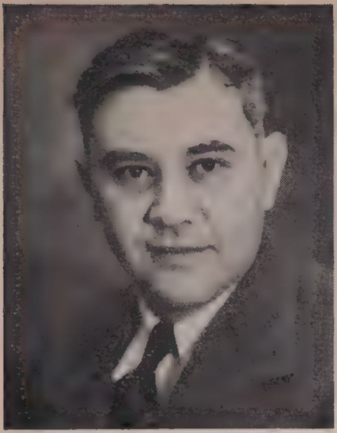
W. S. Bachman (S'32–A'36–M'45) was born at Williamsport, Pa., on October 29, 1908. He received the degree of Electrical Engineer from Cornell University in 1932, and joined the radio-receiver engineering department of the General Electric Company in 1934. His work has included the design and development of loudspeakers, receivers, electromechanical devices, and magnetic circuits.

Mr. Bachman is a member of Tau Beta Pi and Eta Kappa Nu.

❖

HAROLD L. BROUSE

Harold L. Brouse (A'43–M'44) was born on July 16, 1907, at Akron, Ohio. He received the B.S. degree in electrical engineering from Case School of Applied Science, in 1930. From 1930 to 1933 he was a graduate



Madison Cawein

◆

teaching assistant at Rutgers University and received the M.S. degree in 1932.

From 1933 to 1937 he was associated with the B. F. Goodrich Company, as electrical engineer. In 1937 Mr. Brouse joined the Ohio Brass Company, where he was engaged in the development of high-voltage bushings until 1942. Since August, 1942, he has been with the Crosley Corporation as a member of their research staff.

Mr. Brouse is a member of the American Institute of Electrical Engineers, a registered professional engineer of Ohio, and is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

◆

Madison Cawein (M'36–SM'43) was born at Louisville, Kentucky, on March 18, 1904. He received a B.S. degree in physics from the University of Kentucky in 1924, and did graduate work at Cornell University in 1925, and at the University of Kentucky from 1926 to 1928.

In 1925 and 1926 Mr. Cawein was employed by the Westinghouse Lamp Company, in Bloomfield, New Jersey, in the physics laboratory, and later in 1926, was associated with the research department at the Brooklyn Edison Company. From 1928 to 1930 he was an instructor in physics at the University of Kentucky, a post which he left to accept a position as assistant in the astronomical observatory at Princeton University, but accepted a position at Hazeltine instead.

Mr. Cawein was employed by the Hazeltine Corporation from 1930 to 1938, and served as a television consultant for the Andrea Corporation in 1938 and 1939. He joined the staff of the Farnsworth Television and Radio Corporation in 1939, and until 1942 was engaged in television-receiver development with that organization in Marion, Indiana. Since 1942 he has been manager of research at the Fort Wayne, Indiana, plant of the Farnsworth corporation.

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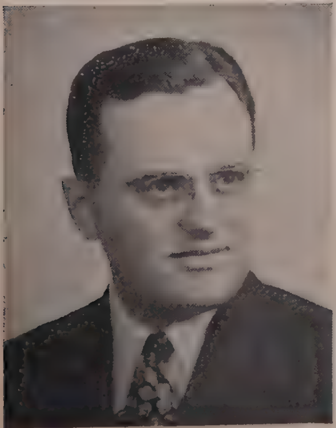


Thomas Studios

Murlan S. Corrington

◆

Murlan S. Corrington was born at Bristol, South Dakota, on May 26, 1913. He received the B.S. degree in electrical engineering in 1934, from the South Dakota School of Mines and Technology, and the M.Sc. degree in 1936, from Ohio State University.

From 1935 to 1937 Mr. Corrington was a graduate assistant in the physics department of Ohio State University. In 1937 he joined the Rochester Institute of Technology, and taught mathematics, mechanics, and related subjects. Since 1942 he has been engaged in mathematical engineering in the advanced-development section of the RCA Victor Division of the Radio Corporation of America. He is the author of "Applied Mathematics for Technical Students," and co-author of "Machine Shop Practice and Instrument Making."

◆

H. G. MacPherson

◆

H. G. MacPherson was born at Victorville, California, on November 2, 1911. He received the A.B. degree in 1932 and the Ph.D. degree in physics in 1937, from the University of California. He has been employed as a research physicist in the research laboratories of the National Carbon Co., Inc., since 1937, engaged in research on the carbon arc and on fundamental properties of carbon.

Myril B. Reed

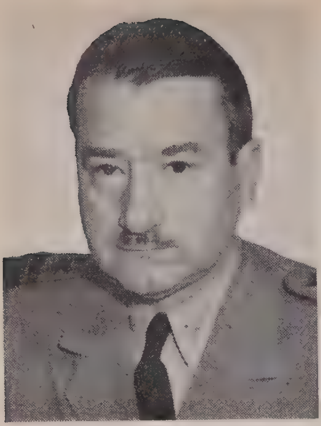
Myril B. Reed (A'41) was born in Woodruff, Utah, on February 14, 1902. He received the B.S. degree in electrical engineering from the University of Colorado in 1926, the M.S. degree in electrical engineering in 1931, and the Ph.D. degree in physics in 1935, both of the latter from the University of Texas. During 1924 and 1925, he worked in the meter department of the Public Service Company of Colorado. For a year after receiving his Bachelor's Degree, he worked as operator in hydro plants of the Utah Power and Light Company. In 1927 Dr. Reed became an instructor in electrical engineering at the University of Texas, and in 1938 he became assistant professor at Illinois Institute of Technology (at that time Armour Institute of Technology) where he is at the present time a professor in electrical engineering, on leave of absence to work with the Bell Telephone Laboratories in New York City. He is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

Dr. Reed published "Fundamentals of Electrical Engineering" in 1938 and has a second book, "Fundamentals of A.C. Circuit Theory," ready for publication. He also has published several papers on circuit theory.

◆

S. A. Schelkunoff (A'40–F'44) received the B.A. and M.A. degrees in mathematics from the State College of Washington in

S. A. Schelkunoff



Walter G. Schindler

◆

1923, and the Ph.D. degree in mathematics from Columbia University in 1928. He was in the engineering department of the Western Electric Company from 1923 to 1925; the Bell Telephone Laboratories from 1925 to 1926; the department of mathematics of the State College of Washington, 1926 to 1929; and Bell Telephone Laboratories, 1929 to date. Dr. Schelkunoff has been engaged in mathematical research, especially in the field of electromagnetic theory.

◆

Walter G. Schindler was born at New Glarus, Wisconsin, on December 10, 1897. He attended St. Johns Military Academy and the University of Wisconsin, and was graduated from the United States Naval Academy in 1921. He also took a graduate course in ordnance at the same institution.

During his career in the Navy, Captain Schindler has specialized in ordnance, serving both at sea and in the Bureau of Ordnance. During the early part of World War II, he served in the Pacific area as gunnery officer and chief of staff for a task force commander. Since 1943 he has served as officer in charge of the Naval Ordnance Laboratory, and has had a leading part in planning the plant now under construction for this laboratory, at White Oak, Maryland.

◆

Kurt Schlesinger

◆

Kurt Schlesinger (A'41) was born on April 20, 1906, in Berlin, Germany. In 1928 he received the engineer's diploma, and in 1929 the degree of Doctor of Applied Physics, both from Technische Hochschule in Berlin, From 1929 to 1930 he was research physicist in Ardenne Research Laboratory, Berlin, and from 1931 to 1937 was chief engineer in the television department of Loewe Radio Company, also in Berlin. In 1938 Dr. Schlesinger became affiliated with the Radio and Cables-Grammont in Paris, France. From 1941 to 1944 he was research engineer for RCA Laboratories. Since that time he has been a member of the research department of the Columbia Broadcasting System, specializing in color television.

◆

Alfred W. Simon

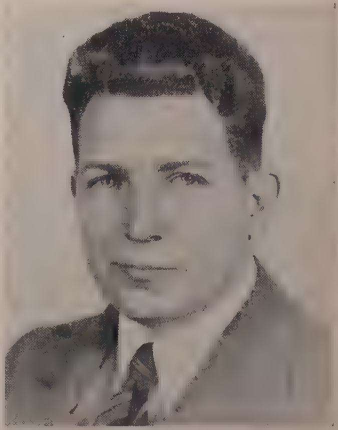
Alfred W. Simon was born in Chicago, Illinois, on September 16, 1897. He studied physics and mathematics at the University of Chicago, receiving the degrees of B.S. in 1921 and Ph.D. in 1925. Subsequently, as a National Research Fellow in Physics at the California Institute of Technology, he carried on researches on electrostatic generators. Entering the industrial field in 1927, he was appointed director of the Cottrell Research Laboratory of the Tennessee Coal, Iron and Railroad Company, a subsidiary of the United States Steel Corporation, and held this position until 1932. From 1935 to 1937 he was employed as research physicist in the radio laboratory of the Stewart Warner Corporation, where he specialized in the design of coils for radio receivers; from 1937 to 1939 he was employed as a geophysicist by the Geophysical Research Corporation and the Stanolind Oil and Gas Company of Tulsa, Oklahoma; and from 1939 to 1941 he was chief engineer of the American Harmonica Company in Chicago. At the outbreak of the war, Dr. Simon transferred to the Naval Ordnance Laboratory of the United States Navy Yard in Washington, D. C., where he stayed until 1943 when he was appointed instructor in mathematics and communication engineering in the Army Specialized Training Program at Washington University in St. Louis. He was employed as chief engineer with John Luellen and Company at Hazel Crest, Illinois from June to December, 1944, and since January 1, 1945, he has been director of research with Edwin I. Guthman and Company, Chicago, Illinois, engaged in research on coils and wire.

◆

L. L. Winter (M'45) was born in Vancouver, Washington, on September 18, 1911. He received the Ph.D. degree in chemistry from the State College of Washington in 1939. Since leaving college he has been employed as a chemist in the research laboratories of the National Carbon Co., Inc., in Cleveland, studying fundamental properties of graphite with particular emphasis on improvement of graphite for electronic applications.

◆

For a photograph and biographical sketch of Charles W. Harrison, Jr., see the June, 1945, issue of the Proceedings; for R. J. Kircher, see the February, 1945, issue of the Proceedings.



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SECTION MEETINGS

Boston

"The Anatomy of Atomic Nuclei," by Robley D. Evans, Massachusetts Institute of Technology; October 19, 1945.

Chicago

"Behavior of Dielectrics Over Wide Ranges of Frequency, Temperature, and Humidity," by R. F. Field, General Radio Company; September 21, 1945.

"Late Developments in Wire Recording," by Hugh Davis, J. P. Seeburg Corporation; September 21, 1945.

Dallas-Ft. Worth

"A Discussion of Quartz Crystals, Their Use and Care," by Merrill Eidson, Eidson's Commercial Crystals and Mountings; August 24, 1945.

Dayton

"Microwave Communications," by Vincent Learned, Sperry Gyroscope Company; October 18, 1945.

Detroit

"Microwave Electronics," by J. A. Hutcheson, Westinghouse Electric Corporation; September 21, 1945.

"Squelch Circuits in Frequency-Modulation and Amplitude-Modulation Recivers," by F. M. Hartz, Detroit Edison Company; October 19, 1945.

Emporium

"The Story of the Proximity Fuze," by W. R. Jones, Sylvania Electric Products, Inc.; October 11, 1945.

Kansas City

"The Behavior of Dielectrics Over Wide Ranges of Frequency, Temperature, and Humidity," by R. F. Field, General Radio Company; September 24, 1945.

"A High-Quality Loudspeaker System of Small Dimensions," by P. W. Klipsch, United States Army; October 9, 1945.

London (Canada)

"Radar in World War II," by K. R. Patrick, RCA Victor; October 5, 1945.

Los Angeles

"Pulse-Position Modulation in Army Communication," by F. J. Altman, United States Army Air Forces, G. M. Snow, and S. A. Voelker, United States Army Signal Corps; October 16, 1945.

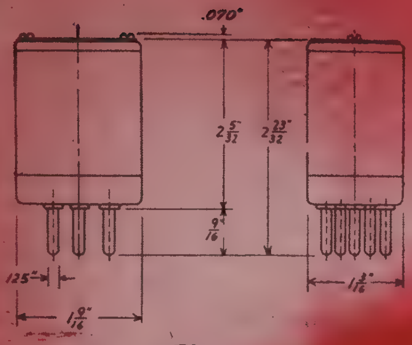
Montreal

"The Application of Radio-Frequency Dielectric Heating to the Dehydration of Penicillin," by G. H. Brown, RCA Laboratories; October 17, 1945.

New York

"Frequency-Modulation Circular Antenna," by M. W. Scheldorf, General Electric Company; July 11, 1945.

(Continued on page 36A)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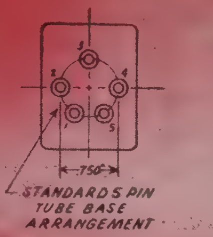
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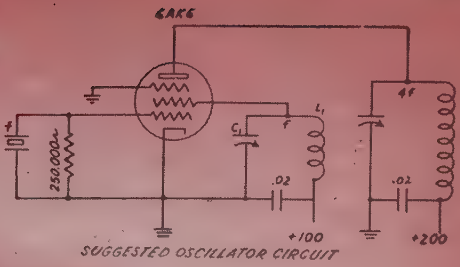
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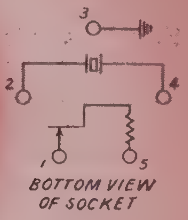
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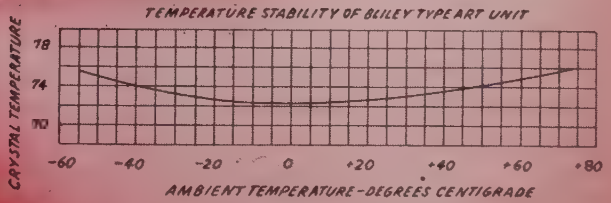
BOTTOM VIEW OF SOCKET



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SECTION MEETINGS

(Continued from page 34A)

"The Application of Cathode-Ray Tubes," by W. J. Knoop, Jr., Allen B. DuMont Laboratories; July 11, 1945.

"Radar in the United States Army," by R. B. Colton, United States Army Air Forces; September 5, 1945.

"The Gun Director's Conquest of the Robot Bomb," by C. A. Lovell, Bell Telephone Laboratories; September 19, 1945.

"Ratio Detectors for Frequency-Modulation Receivers," by S. M. Seeley, RCA Laboratories; October 3, 1945.

PHILADELPHIA

"The Federal Communications Commission—Its Consideration of Technical Matters Relating to the Various Communication Services," by V. R. Simpson, Federal Communications Commission; October 4, 1945.

PITTSBURGH

"The Radio Sonde and Associated Meteorological Equipment," by L. D. Whitelock, Bureau of Ships; October 8, 1945.

PORTLAND

Discussion on Frequency Modulation; October 1, 1945.

ST. LOUIS

"A High-Quality Loudspeaker System of Small Dimensions," by P. W. Klipsch, United States Army; October 24, 1945.

SAN DIEGO

"Radio-Frequency Curing of Conōlon," by B. D. Abramis, Consolidated-Vultee Aircraft Corporation; October 9, 1945.

TWIN CITIES

"The Behavior of Dielectrics Over Wide Ranges of Frequency, Temperature, and Humidity," by R. F. Field, General Radio Company; September 25, 1945.

WASHINGTON

"Very-High-Frequency Train Communication Systems," by J. W. Hammond, Bendix Aviation Corporation; October 9, 1945.

"Inductive Train Communication," by L. O. Grondahl, Union Switch and Signal Company; October 9, 1945.

WILLIAMSPORT

"New Amplitude-Modulation, Frequency-Modulation and Television Studios at WHAM-WHFM," by K. J. Gardner, Radio Station WHAM-WHFM; October 3, 1945.

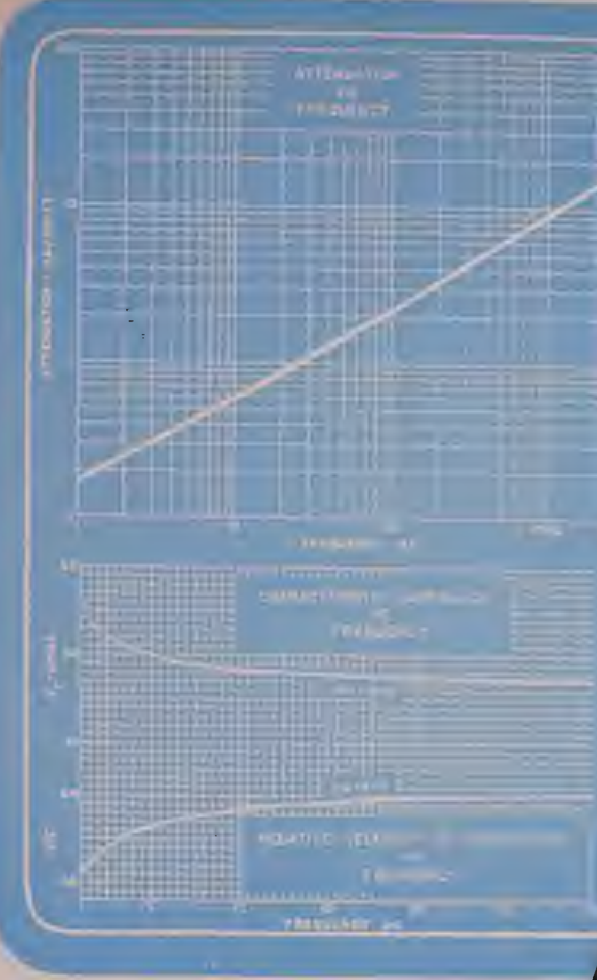
SUBSECTIONS

COLUMBUS

"A Fundamental Pulse Radar System," by D. A. Newberry, United States Navy; September 14, 1945.

"Industrial Electronic Control," by K. H. Keller, General Electric Company; October 12, 1945.

(Continued on page 38A)



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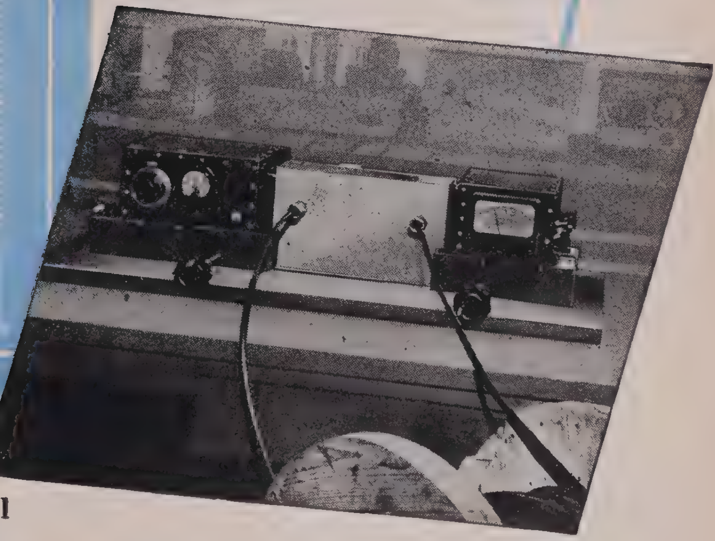

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SECTION MEETINGS

(Continued from page 36A)

MILWAUKEE

"Behavior of Dielectrics Over Wide Ranges of Frequency, Temperature, and Humidity," by R. F. Field, General Radio Company; September 21, 1945.

PRINCETON

"Radar Sets and How They Grew," by L. A. Turner, Radiation Laboratory, Massachusetts Institute of Technology; October 12, 1945.

SOUTH BEND

"The Behavior of Dielectrics Over Wide Ranges of Frequency Temperature and Humidity," by R. F. Field, General Radio Company; September 27, 1945.

MEMBERSHIP

The following transfers and admissions were approved on November 7, 1945:

Transfer to Senior Member

Beach, A. R., 307 Poplar Dr., Falls Church, Va.
Briggs, T. H., Superior Tube Co., Norristown, Pa.
Burroughs, H. A., 2124 Key Blvd., Arlington, Va.
Carroll, M. J., 589 E. Illinois St., Chicago 11, Ill.
Dondanville, R. V., 3909 Johnson Ave., Western Springs, Ill.
East, L. A. W., 204 Hospital St., Montreal 1, P. Q., Canada
Freundlich, M. M., 9 W. 87 St., New York 24, N. Y.
Gardiner, P. C., 210 James St., Scotia, N. Y.
Hastings, T. M., Jr., 530 Commonwealth Ave., Boston, Mass.
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(Continued on page 40A)

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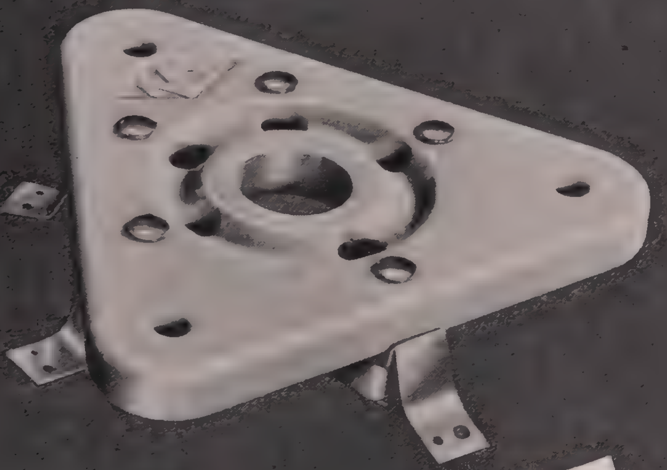

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(Continued on page 42A)

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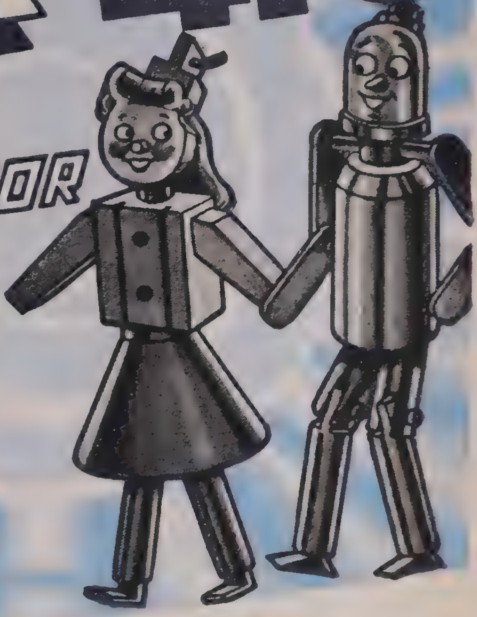

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(Continued on page 44A)

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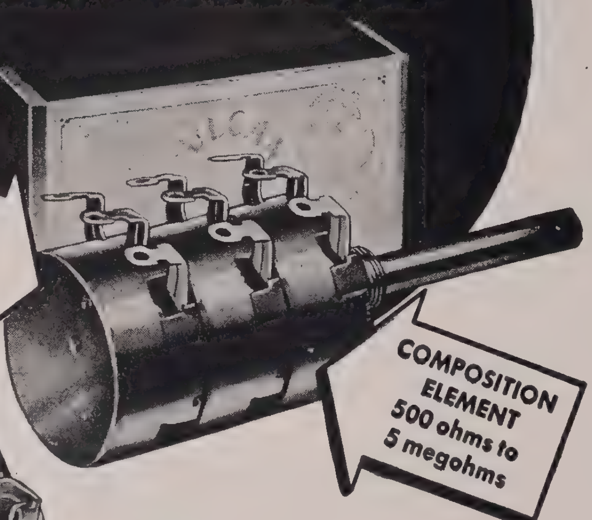

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(*Continued on page 46A*)

Down to fundamentals...

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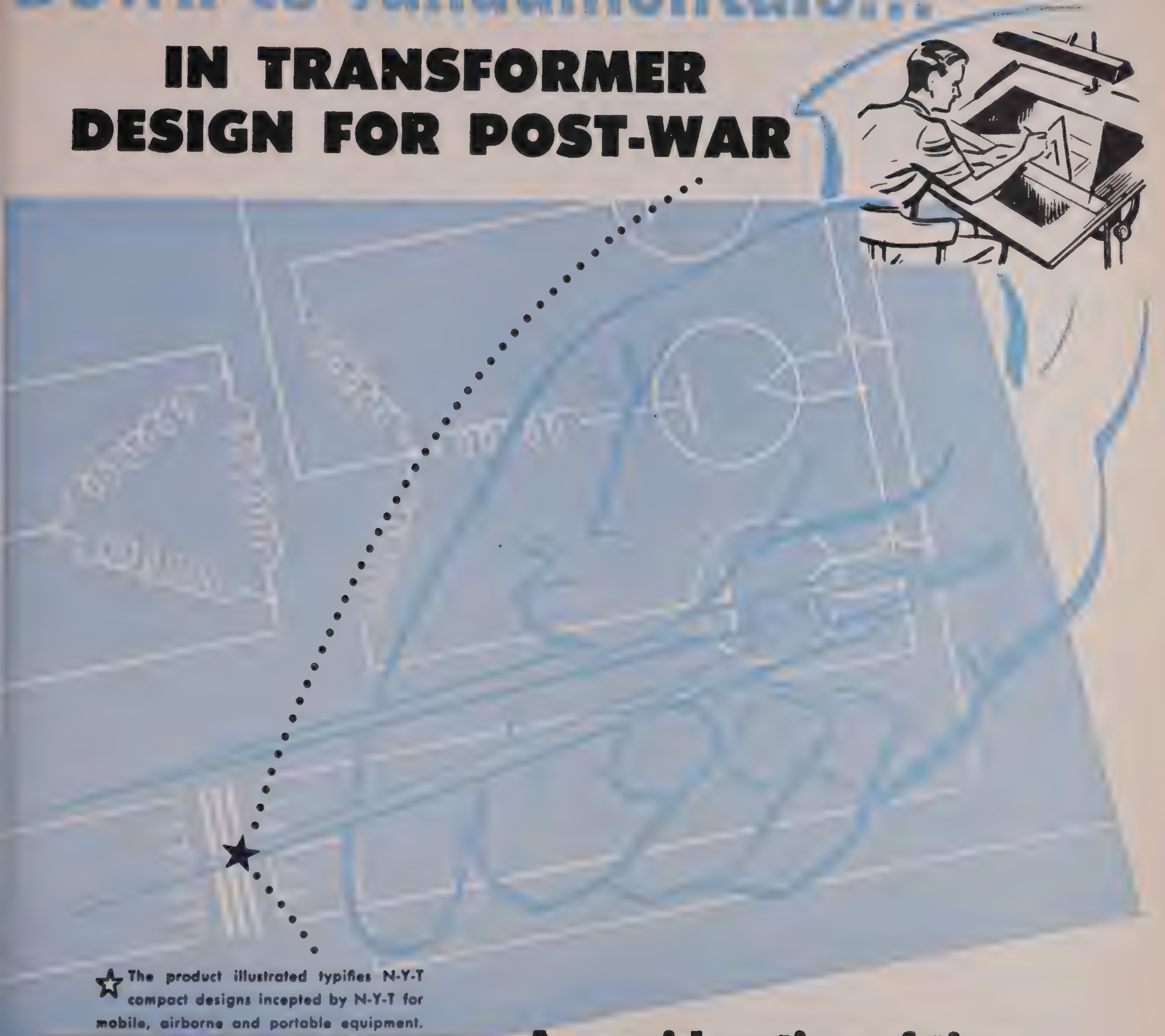

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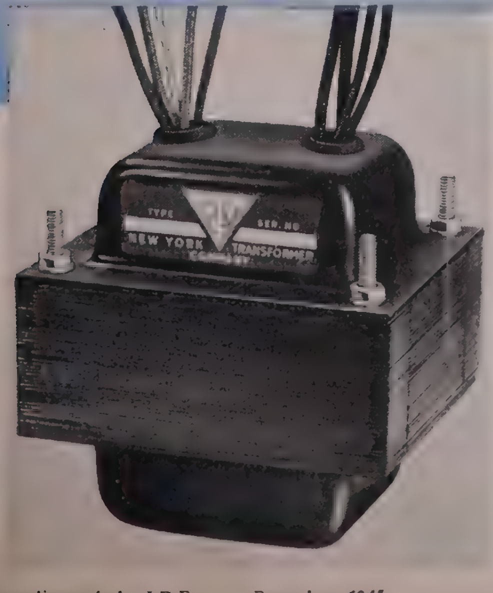

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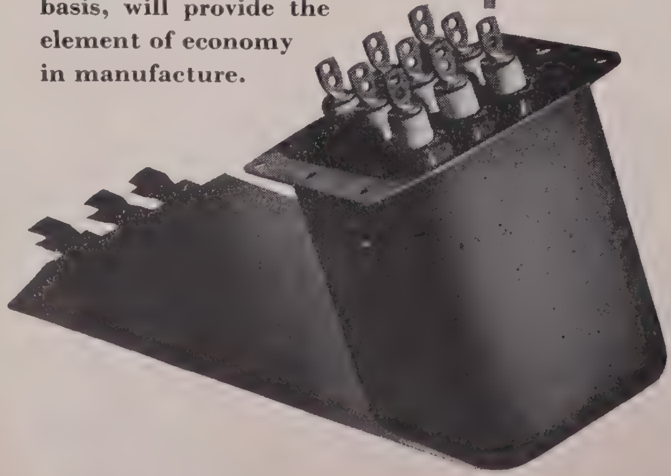

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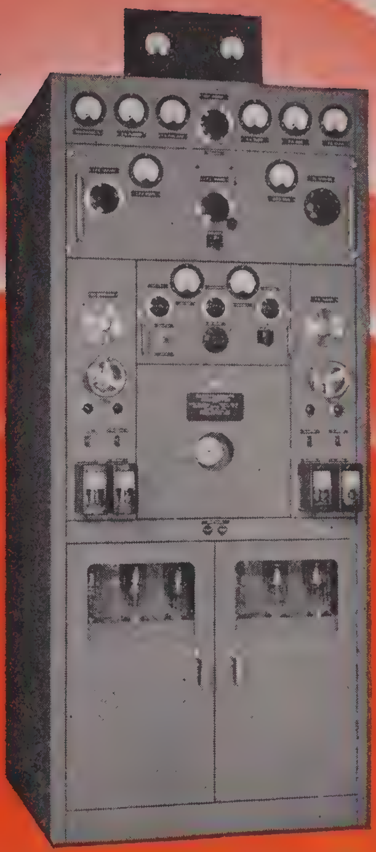

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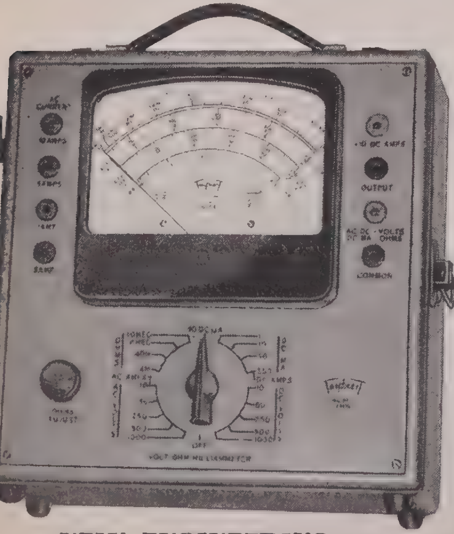

MODEL 2405

Volt · Ohm Milliammeter

25,000 OHMS PER VOLT D.C.

STANDARDS ARE SET BY TRIPLETT

SPECIFICATIONS

NEW "SQUARE LINE" metal case, attractive tan "hammered" baked-on enamel, brown trim.

■ PLUG-IN RECTIFIER — replacement in case of overloading is as simple as changing radio tube.

■ READABILITY—the most readable of all Volt-Ohm-Milliammeter scales—5.6 inches long at top arc.

NEW ENGINEERING
NEW DESIGN • NEW RANGES
30 RANGES

Voltage: 5 D.C. 0-10-50-250-500-1000 at 25000 ohms per volt.
5 A.C. 0-10-50-250-500-1000 at 1000 ohms per volt.
Current: 4 A.C. 0-.5-1-5-10 amp.
6 D.C. 0-50 microamperes—0-1-10-50-250 milliamperes—0-10 amperes.
4 Resistance 0-4000-40,000 ohms—4-40 megohms
6 Decibel -10 to +15, +29, +43, +49, +55
Output Condenser in series with A.C. volt ranges

Model 2400 is similar but has D. C. volts Ranges at 5000 ohms per volt.

Write for complete description

Triplett

ELECTRICAL INSTRUMENT CO.

BLUFFTON ■ OHIO

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(Continued on page 58A)

LARGE OR SMALL

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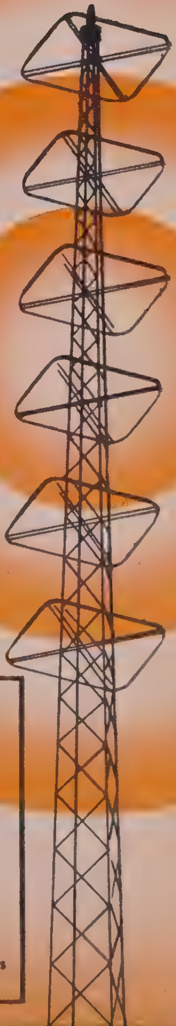
SPECIFY
AMERTRAN
Components
in your
NEW
TRANSMITTER

Rating: 74DB, Class B, Modulation Transformer .03-15KC; 12 KV D.C. Operating 51 K.V. Hi-potential.

There is an AmerTran Transformer for practically every transformer symbol in all electronic circuits. They range from units to fit in a microphone to the huge modulation and plate transformers used by large broadcasting stations. Whatever their size, they have one thing in common—the quality construction that has made AmerTran the leader in transformer manufacture for forty-four years. When writing for Bulletin state product in which you are interested.



Modulation Transformers and Reactors
Oil or Abestol Immersed Plate Transformers
W Type Transmitter Transformers
Transtat A. C. Voltage Regulators
(Manual, Motor and Automatic Operation)
Radio components
Wave filters
Rectifiers
Transformers for Electronic Applications
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MANUFACTURING SINCE 1901 AT NEWARK N. J.

Pioneer Manufacturers of Transformers, Reactors and Rectifiers for Electronics and Power Transmission

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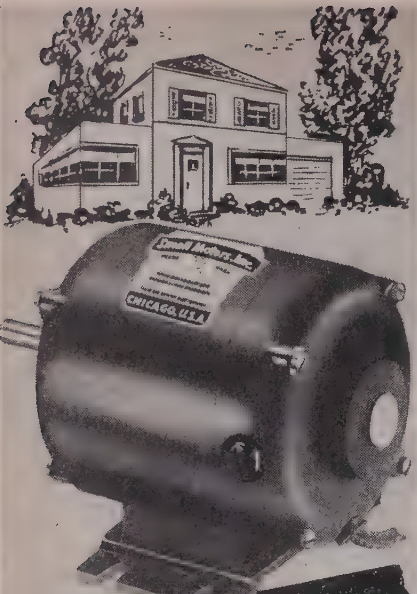
AMERICAN TRANSFORMER COMPANY
178 EMMET STREET, NEWARK 5, N. J.



Quiet • POWERFUL SM Fractional H. P. Motor

The Home Power Servant also handles many other jobs efficiently, dependably

For quiet operation, dependable performance, long life, maximum power per ounce of weight and per inch of space, use SM Fractional H.P. Motors. Models from 1/10th to 1/200th H.P. Speeds of 3,000 to 20,000 R.P.M. Voltage from 6 to 220 AC-DC Large volume production to your exact specifications.

Small Motors, Inc.

DEPT. 31
1308 ELSTON AVE. • CHICAGO 22

Manufacturers of special small universal, fractional H. P. motors, dynamotors, shaded pole motors, heater motors, generators.

Design, Engineering, Production

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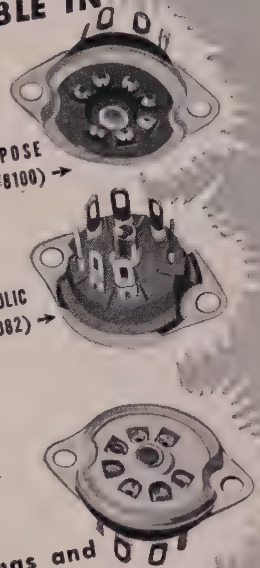
(Continued on page 60A)

EBY SADDLE TYPE MINIATURE TUBE SOCKETS

AVAILABLE IN

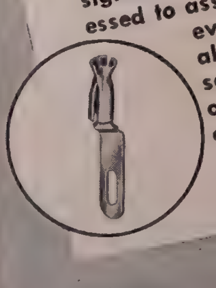


GENERAL PURPOSE PHENOLIC (EBY #8100) →
LOW LOSS PHENOLIC (EBY #8082) →
STEATITE (EBY #8083) →

The castings and beryllium copper contacts are identical with JAN S-28 Types SO 10M and SO 10C except that the shield base is replaced with a saddle. These EBY sockets meet the need for quality replacement of sockets of the saddle type. Write today for prices and samples.

LONG LIFE CONTACTS

The self-aligning beryllium copper contacts have been especially designed and Micro-processed to assure constant, even pressure on all parts of the socket pin without fatigue in contacts after continuous use.



IF IT'S IN A CIRCUIT . . . EBY COMPONENTS AND SERVICES WILL HELP YOU DO IT BETTER

HUGH H. EBY INCORPORATED
18 W. CHELTEN AVE.
PHILADELPHIA, PA.

GROUPED about the widely acclaimed Two-million-Volt Precision X-ray Tube are other Machlett tubes for medical, industrial and radio purposes. In each of these tubes are incorporated the inherent skills employed by Machlett in the development of this unique tube. They are your assurance of long life, ruggedness and dependability in whatever field they are used. Machlett Laboratories, Inc., Springdale, Connecticut.

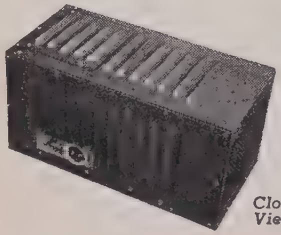
MACHLETT

APPLIES TO RADIO AND INDUSTRIAL USES
ITS 48 YEARS OF ELECTRON-TUBE EXPERIENCE

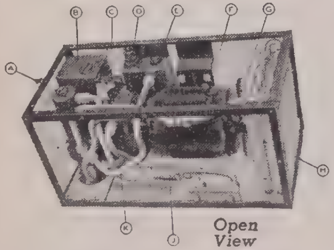
EASTERN HEAT DISSIPATING UNIT

The Eastern Heat Dissipating Unit is used in connection with television, radar, short wave radio communications, high pressure mercury lamps, X-Ray tubes, induction heating units, and many other applications. It was developed for military requirements in conjunction with radar and electronic tube cooling problems. Units were designed in various sizes and capacities, some with the close heat control range of 2 degrees C. Used successfully for ground, water and airborne service, they combine rugged construction, compactness and light weight.

The model illustrated will dissipate up to 1200 watts with a constant controlled temperature, irrespective of surrounding temperatures, within 2 degrees C. It is complete with Thermostat control, Thermostatic valves and flow switch. Eastern has built airborne units of much smaller sizes and industrial units of much larger sizes and capacities. The specifications for the unit shown are: SIZE: 16" x 7½" x 7½"; METALS: Steel, Bronze, or Aluminum. Other models can be designed to dissipate up to 5000 watts.



Closed View



Open View

A. Thermo flow-regulator	*F. Motor*
B. Reservoir	*G. Fan*
C. Flow Switch	*H. Radiator*
D. Thermostat	*J. Filter*
E. Auxiliary Heater	*K.Pump*

Eastern's experience in solving heat control problems, especially where compactness and light weight are necessary, makes them the logical people with whom to discuss heat control applications. If you are designing or planning to build equipment that calls for heat dissipation or the close control of operating temperatures, Eastern will design and build the entire unit for you to meet your specific requirements.

An inquiry about your heat dissipating needs will not obligate you in the slightest.

A large part of Eastern's business is the designing and building of special pumps, in quantities ranging from 25 to several thousand for the aviation, electronic, chemical, machine and other special fields. Eastern builds over 600 models, both centrifugal and positive pressure, ranging in size from 1/100 H.P. to ¾ H.P. as standard units.

Eastern Engineering Co.

86 FOX STREET, NEW HAVEN 6, CONN.

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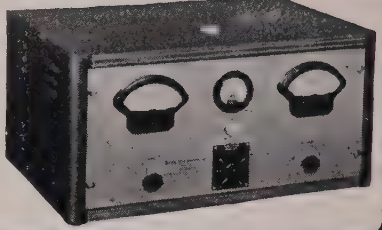
(Continued on page 62A)

Engineered and built by

Doolittle



FM and AM
FREQUENCY MONITORS

Direct reading. No charts or complicated calculations necessary. Models available for 110 volt A.C. or battery operated portable use. Meet FCC requirements.

DISTORTION METER

Direct reading device which indicates as a percentage of the fundamental frequency, the square root of the sum of the squares of the harmonic components. It is used for audio frequency measurements in any audio device in the usual range of voice or musical notes from 150 to 15,000 cycles.

• Utilize the many advantages of these units now. They are sturdily built, self-contained, moderately priced. Remember . . . equipment pioneered by DOOLITTLE years ago, *still serves efficiently today!*

SEND FOR FULL DETAILS

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BUILDERS OF PRECISION RADIO EQUIPMENT

IRON SLEEVE CORES

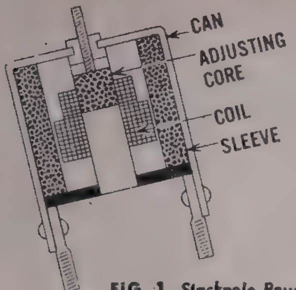


FIG. 1. *Stackpole Powdered Iron Sleeve and Core used for Diode Transformer (I-F); Antenna, Oscillator, or Filter Coils, etc.*

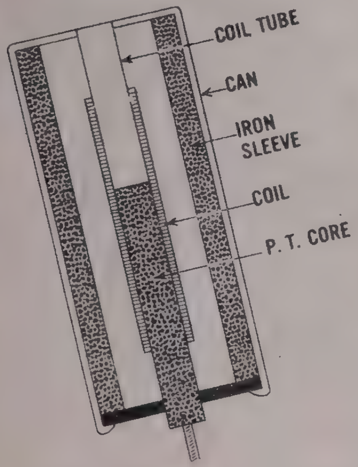


FIG. 2. *Grade SK1 core and powdered iron sleeve (.790 O. D. x 1½" long) used with permeability tuning in auto radio receiver.*

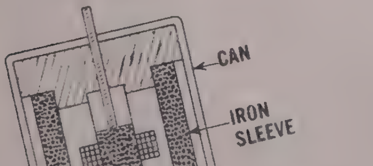


FIG. 3. *Two Stackpole cores and powdered iron sleeve used in a double tuned I-F transformer application.*

The Modern Answer to Better Coils in Less Space

BY USE of Stackpole Sleeve Cores, much smaller cans *of any material* may be used to provide Q that is equal to or better than that of conventional coils and cans. Thus they pave the way to an exceptionally high order of tuning unit efficiency in greatly reduced size. A few of many design possibilities are indicated in the accompanying sketches.

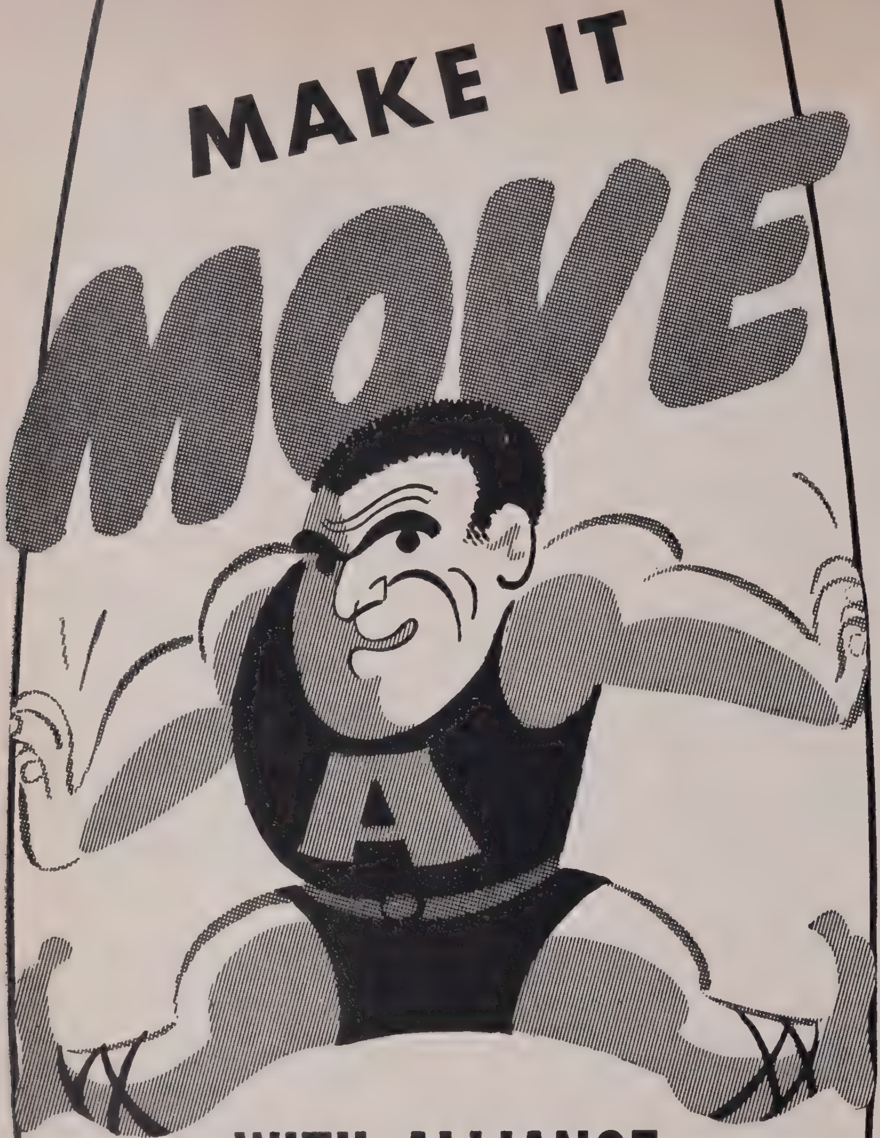
Beside supplying additional electrostatic and electromagnetic protection over that provided by the can alone, sleeve cores result in making the can itself smaller, less critical and less costly. Inexpensive die cast lead cans, for instance, may be used instead of aluminum. In some cases, it may not even be necessary to use a can.

STACKPOLE CARBON COMPANY, ST. MARYS, PA.

EXPORT: Stackpole Carbon Co., 254 W. 34th St., New York 1, N. Y., U. S. A.

STACKPOLE

ELECTRICAL BRUSHES AND CONTACTS *(All carbon, graphite, metal and composition types)* **• RARE METAL CONTACTS • WELDING CARBONS • BRAZING TIPS AND BLOCKS • PACKING, PISTON, AND SEAL RINGS • CARBON REGULATOR DISCS • MOLDED METAL COMPONENTS, ETC.**



WITH ALLIANCE

Thousands of little motors leave the high speed production lines of Alliance . . . every day. Where will they go? Wherever you need compact, condensed power sources . . . power controlled from a central point.

It will pay you to investigate Alliance motors . . . they will make your product *MOVE* . . . bring it to life.

Long experience in the advanced design and manufacture of miniature electric motors makes for high quantity, low cost production!

WHEN YOU DESIGN—KEEP
alliance
IN MIND

ALLIANCE MANUFACTURING CO. • ALLIANCE, OHIO

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Thornhill, J. R., 81 Balmoral Ave., Toronto, Ont., Canada
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Williams, J. M., 163 N. Euclid Ave., Westfield, N. J.
Wilson, G. L., The Clarendon Laboratory, Parks Rd., Oxford, England
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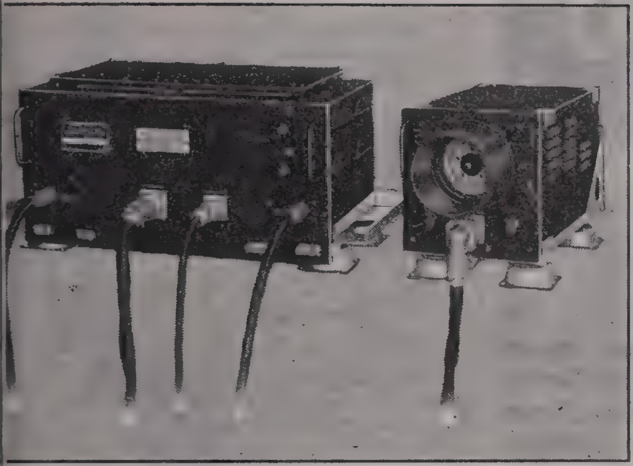
Woodard, T. C., 1802½ N. Western Ave., Los Angeles 27, Calif.
Worster, J. R., 1511 Lincoln Ave., New Hyde Park, N. Y.
Zimmer, E. C., 52 Union St., Hamburg, N. Y.



RCA Radio Altimeters are used on all the global routes of the Air Transport Command

FOR the past three years, the Air Transport Command has supplied the American Air Forces in China by flying supplies over the 20,000-foot peaks of the snow-clad Himalaya Mountains. This flying route with its 100-mile winds, severe downdrafts and cloud-hidden peaks is one of the most dangerous in the world—yet as high as three hundred trips a day must be made over it.

One of the instruments that help to maintain this ATC schedule is the high-altitude model of the RCA family of Radio Altimeters. This altimeter indicates the plane's clearance over cloud-hidden peaks and tells the pilots of the Air Transport Command when it is safe to let down through the overcast.

Thus, once again, RCA Research has made a major contribution to the safety of flying—a contribution that will be equally important in postwar commercial airline service. Remember, "When you think of Radio Altimeters, think of RCA."

Buy More War Bonds

RCA

RADIO CORPORATION OF AMERICA

RCA VICTOR DIVISION • CAMDEN, N. J.

1,001 USES

Condensed Power for Years of Service

VERSATILITY and dependability were paramount when *Alliance* designed these efficient motors — *Multum in Parvo!* . . . They are ideal for operating fans, movie projectors, light home appliances, toys, switches, motion displays, control systems and many other applications . . . providing economical condensed power for years of service.

Alliance Precision

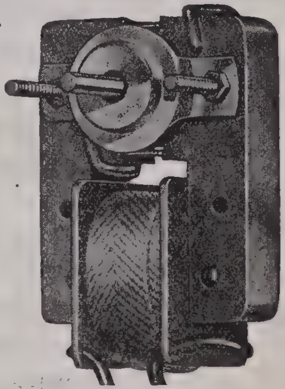
Our long established standards of precision manufacturing from highest grade materials are strictly adhered to in these models to insure long life without breakdowns.

EFFICIENT

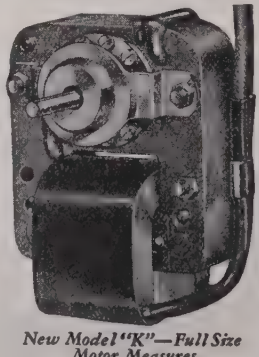
Both the new Model "K" Motor and the Model "MS" are the shaded pole induction type — the last word in efficient small motor design. They can be produced in all standard voltages and frequencies with actual measured power outputs ranging upwards to 1/100 H. P. . . Alliance motors also can be furnished, in quantity, with variations to adapt them to specific applications.

DEPENDABLE

Both these models uphold the *Alliance* reputation for all 'round dependability. In the busy post-war period, there will be many "spots" where these Miniature Power Plants will fit requirements . . . Write now for further information.



Model "MS" — Full Size Motor Measures 1¾" x 2 x 3⅛"



New Model "K" — Full Size Motor Measures 2⅛" x 2⅜" x 3$\frac{1}{16}$"

Remember Alliance!
—YOUR ALLY IN WAR AS IN PEACE

ALLIANCE
MANUFACTURING CO.
ALLIANCE . OHIO

SECTION MEETINGS

Boston

"Microwave-Oscillation Generators Using Velocity-Modulated Electron Beams," by E. U. Condon, Westinghouse Electric Corp.; April 27, 1945.

Buffalo-Niagara

"Radio-Receiver Design and Measurements," by Byron Atwood, Colonial Radio Corp.; April 18, 1945.

Cedar Rapids

"Management Problems of a Manufacturer of Engineering Products," by A. E. Thiessen, General Radio Co.; April 19, 1945.

"Unusual Tube Circuits," by E. L. Kent, C. G. Conn, Ltd.; May 18, 1945.

Chicago

"Current Problems in Radio Engineering and Manufacturing," by Paul Galvin, Galvin Manufacturing Corp.; May 18, 1945.

Cincinnati

"Pulse Modulation," by G. F. Leydorf, Crosley Corp.; April 17, 1945.

Dallas-Fort Worth

"Aircraft Instrument Landing," by H. A. Smith, Civil Aeronautics Administration; April 6, 1945.

Dayton

"Behind the Instrument Dial," by L. F. Parachini, Weston Electrical Instrument Co.; May 17, 1945.

Emporium

"Telephone Service for Vehicles and Vessels in Urban Areas," by R. T. Griffith, Bell Telephone Company of Pennsylvania; May 16, 1945.

London

Student Night; April 13, 1945.

"The Cathode-Ray Tube," by B. W. Graham, Sparton of Canada; April 26, 1945.

Montreal

"Ultra-High-Frequency Measurements" by D. B. Sinclair, General Radio Company; March 28, 1945.

"Mathematical Tools," by Faivel Panter, Canadian Marconi Company; April 11, 1945.

"The Application of Radio to Railroad Signaling Systems," by W. S. Halstead, Halstead Traffic Communications Corp.; May 9, 1945.

Election of Officers; May 9, 1945.

Inspection Tour of Traffic-Control Centre of the Canadian National Railways Central Station Terminal; May 23, 1945.

"Details of Operation of Traffic-Control Equipment," by E. J. Agnew, Union Switch and Signal Company of New York; May 23, 1945.

Ottawa

"Ionosphere Focusing," by Wilbur Smith, Department of Transport, Canadian Government; February 27, 1945.

"The Magic of the Spectrum," by J. W. Bateman, Canadian General Electric Co.; April 17, 1945.

"The Application of Radio to Railroad Signaling Systems," by William Halstead, Halstead Traffic Communications Corp.; May 22, 1945.

Election of Officers; May 22, 1945.

Philadelphia

"Wartime Developments in Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; May 3, 1945.

Election of Officers; May 3, 1945.

(Continued on page 38A)

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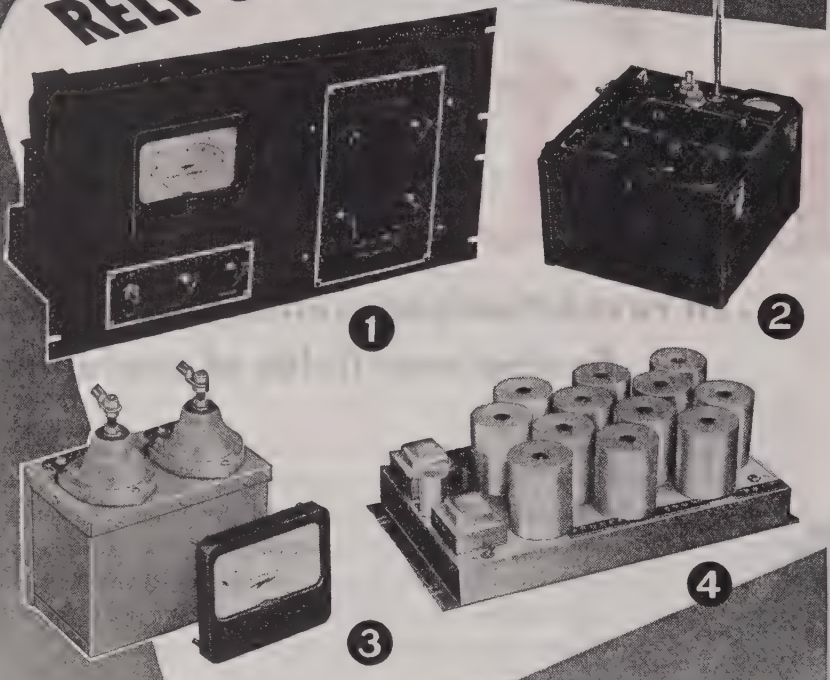

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PORTLAND

"Wartime Developments in Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; April 17, 1945.

ROCHESTER

"Multiplex Radio Link Across Chesapeake Bay," by N. F. Schlaack and C. R. Burrows, Bell Telephone Laboratories, Inc.; May 17, 1945.

SAN DIEGO

"Wartime Developments in Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; April 23, 1945.

"Television," by C. G. Pierce, General Electric Co.; May 21, 1945.

SEATTLE

"Industrial Electronic Development at Schenectady," by H. L. Palmer, General Electric Co.; May 18, 1945.

TORONTO

"Ultra-High-Frequency Measurements," by D. B. Sinclair, General Radio Co.; March 26, 1945.

"Quartz-Crystal Production," by L. W. Elliott, Canadian Marconi Co.; April 9, 1945.

"Signal Operations in the Field," by J. E. Andrews; 1st Divisional Signals, R.C.C.S.; April 9, 1945.

TWIN CITIES

"Television in Practice," by B. R. Hilker, Station KSTP; February 13, 1945.

"Minneapolis-Honeywell Electronic-Tube Regulator System for Aircraft," by Alfred Upton, Minneapolis-Honeywell Regulator Co.; May 8, 1945.

Election of Officers May 8, 1945.

WILLIAMSPORT

"Radio Plans for Postwar Private and Commercial Aviation," by T. B. Bourne, Civil Aeronautics Authority; May 2, 1945.

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Fink, D. G., 1608 N. Queen St., Arlington, Va.
Gilbert, W. E., 1020 E. Haines St., Philadelphia 38, Pa.
Gray, T. S., Electrical Engineering Dept., Massachusetts Institute of Technology, Cambridge 39, Mass.
Herber, J. C., Bell Telephone Laboratories, Inc., 463 West St., New York 14, N. Y.
Hickman, R. W., Cruft Laboratory, Harvard University, Cambridge 38, Mass.
Kuper, J. B. H., 17 Traill St., Cambridge 38, Mass.
Lemmon, W. S., 590 Madison Ave., New York 22, N. Y.
Litton, C. V., Box 749, Redwood City, Calif. (as of November 1, 1944)

(Continued on page 41A)

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(Continued from page 38A)

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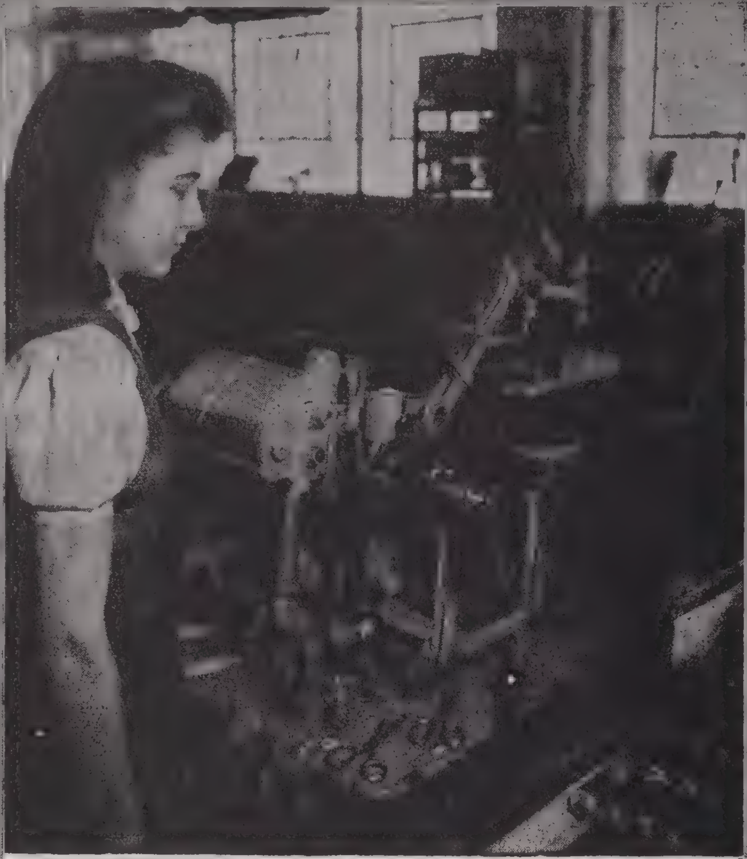

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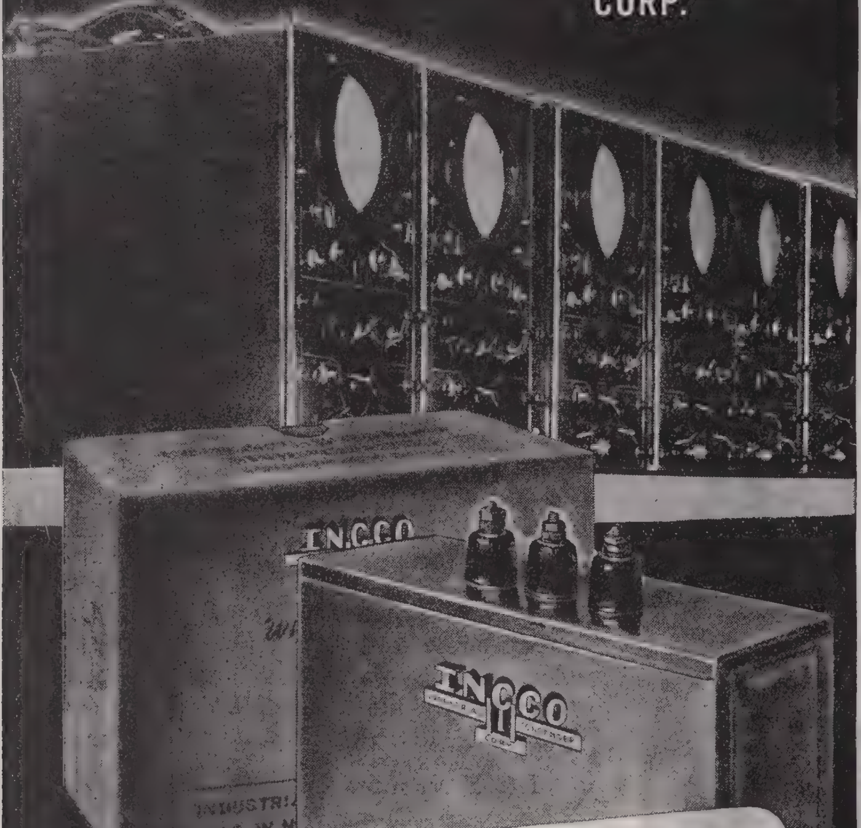

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Thomas, L. W., 3602 S. Dakota Ave., N.E., Washington 18, D. C.
Trent, R. L., 463 West St., New York 14, N. Y.
Wagner, F. R., 4111 Walnut St., Kansas City 2, Mo.
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Bailey, T. J., Box 62, Elm Grove, Wis.
Barlow, A. D., Beverly, West Va.
Bearden, J. A., Radiation Laboratory, Johns Hopkins University, Homewood Campus, 34 and Charles Sts., Baltimore, Md.
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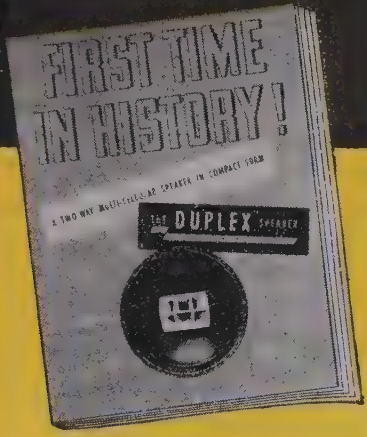

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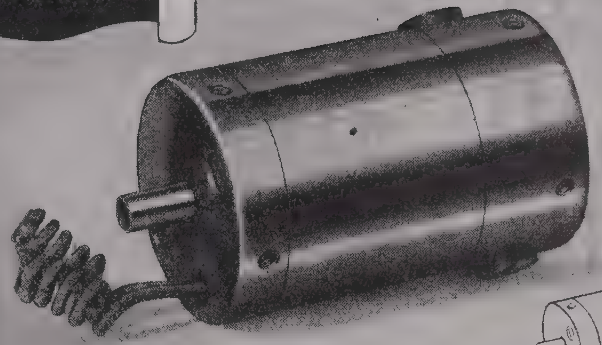

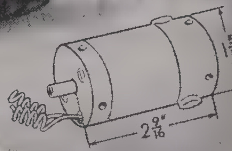

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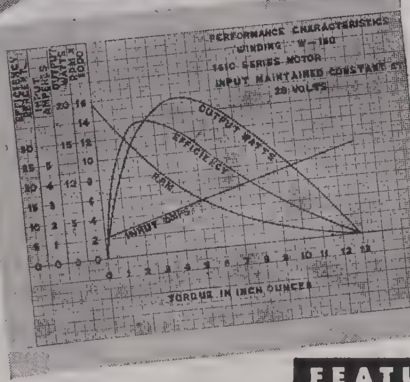

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1600 FRAME MOTORS		Series	Shunt
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Watts Output, Con.	(max.)		5
Torque at 8500 RPM	(in. oz.)	3	
Torque at 5800 RPM	(in. oz.)	4.5	1
Lock Torque	(in. oz.)	12	3
Volts Input	(min.)	5	5
Volts Input	(max.)	32	32
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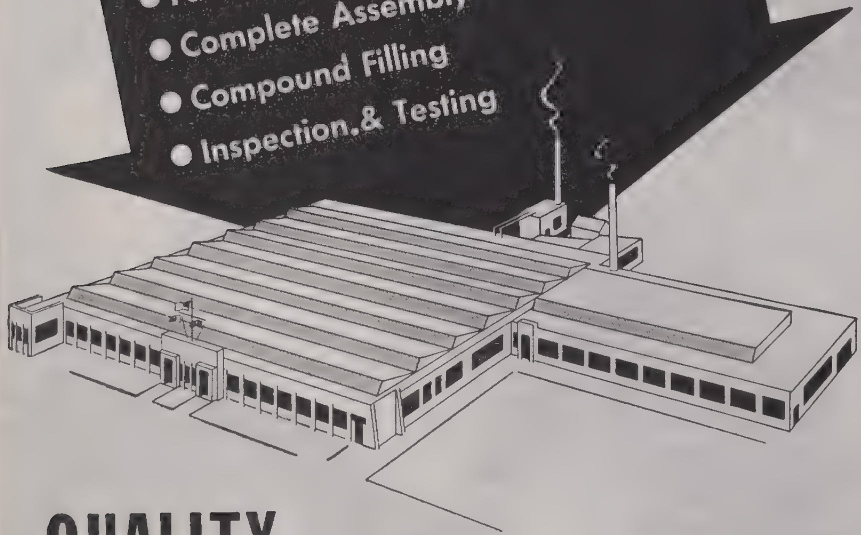

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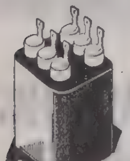

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(Continued on page 69A)

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(*Continued from page 48A*)

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INDEX

Section Meetings 36A
Membership 38A
Positions Open 50A

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Acme Electric & Mfg. Co. 60A
Aireon Mfg. Corp. 1A
Airplane & Marine Instruments, Inc. 50A
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Allied Radio Corp. 46A
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American Transformer Co. 21A
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Andrew Company 38A
Arnold Engineering Co. 11A
Aviometer Corp. 8A

Barker & Williamson 57A
Blaw-Knox Company 41A
Boonton Radio Corp. 67A

Centralab 10A
Chicago Transformer 48A
Clarostat Mfg. Co. 58A
Sigmund Cohn & Co. 67A
Collins Radio Co. 15A
Communications Co., Inc. 69A
Continental-Diamond Fibre Co. 26A
Cornell-Dubilier Electric Corp. Cover III
Corning Glass Works 17A
Coto-Coil Company, Inc. 42A

Daven Company 13A
De-Jur-Amsco Corp. 14A
Tobe Deutschmann Corp. 71A
Allen B. DuMont Labs. 16A

Hugh H. Eby, Inc. 64A
Eicor Inc. 45A
Stanley D. Eilenberger 69A
Eitel-McCullough, Inc. 72A
Electrical Reactance Corp. 62A
Electronic Engineering Co. 54A
Electronic Mechanics, Inc. 23A

Federal Tel. & Radio Corp. 3A, 51A, 68A
Freed Radio Corp. 52A
Freed Transformer Co. 66A

General Electric Co. 19A
General Instrument Corp. 49A
General Radio Co. Cover IV
Guardian Electric Mfg. Co. 29A

Hallicrafters Co. 6A
Harco Company, Inc. 70A
Harvey Radio Labs., Inc. 33A
Hazeltine Corp. 50A
Heintz & Kaufman, Ltd. 20A
Hewlett-Packard Co. 24A
Hytron Radio & Electronics Corp. 25A

Industrial & Commercial Electronics 43A
Industrial Condenser Corp. 44A
International Tel. & Tel. Corp. 3A, 51A, 68A

Jennings Radio Mfg. Co. 47A
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Murray Hill Books, Inc., Tech. Div. 63A

National Carbon Co., Inc. 32A
National Company 31A

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Permoflux Corp. 64A
Premax Products 66A
Press Wireless, Inc. 52A

Radio Corp. of America Facing 32A, Facing 32D, 34A, 35A
Rauland Corp. 70A
Raytheon Mfg. Co. 52A
Remler Co., Ltd. 18A

Santay Corp. 39A
Shallcross Mfg. Co. 51A
Skydyne, Inc. 54A
Sprague Electric Co. 12A
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CINCINNATI

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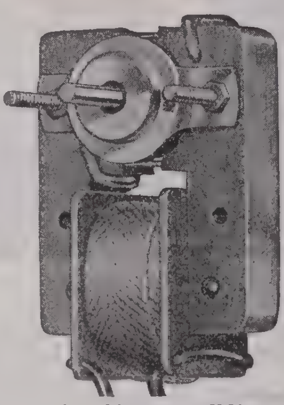

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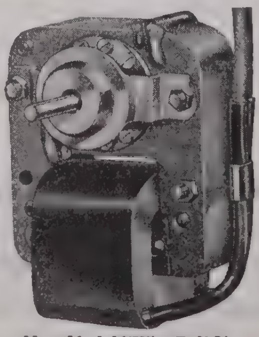

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Section Meetings

(Continued from page 34A)

"Wide-Range Tuned Circuits for High Frequencies," by A. Petersen, Bell Telephone Laboratories, Inc.; March 16, 1945.

"The Servo Problem as a Transmission Problem," by E. B. Ferrell, Bell Telephone Laboratories, Inc.; March 16, 1945.

"The Vibratron," by Frank Rieber, Rieber Electronic Laboratories; March 28, 1945.

"Wartime Developments in Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; April 4, 1945.

"Precision Methods of Electronic Timing," by J. T. Potter, Potter Instruments Company; April 18, 1945.

"Present and Postwar Aviation-Communication Systems," by D. W. Rentzel, Aeronautical Radio, Inc.; April 26, 1945.

"Some Plane Common Sense About Television," by P. I. Merryman, National Broadcasting Company; May 2, 1945.

"Models of the Motion of Electron and Electro- and Magneto-Static Fields," by J. A. Rajchman, RCA Laboratories; May 11, 1945.

"Field Experiences with the Very-High-Frequency Radio-Relay Equipment in Normandy," by A. H. Waite, Jr., Coles Signal Laboratory; May 11, 1945.

"Frequency Allocations for Television," by G. P. Adair, Federal Communications Commission; June 15, 1945.

"Reflective Optics as Applied to Television," by I. G. Maloff, RCA Victor Division; June 15, 1945.

"Refractive Optics in Television Projection," by Konstantin Pestracov, Bausch and Lomb Optical Company; June 15, 1945.

"Direct Viewing Large Image with a Twenty-Inch Tube," by A. B. DuMont, Allen B. DuMont Laboratories, Inc.; June 15, 1945.

"Pros and Cons of Direct Viewing versus Projection Television," by D. D. Israel, Emerson Radio and Phonograph Corporation; June 15, 1945.

"Color-Television Receivers," by P. C. Goldmark, Columbia Broadcasting System; June 15, 1945.

ROCHESTER

"Wartime Developments in Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; May 31, 1945.

ST. LOUIS

"Electronic Heating," by W. D. Wenger, RCA Victor Division; April 26, 1945.

"A Graphical Analysis of Frequency Modulation," by S. H. Van Wambeck, Washington University; May 29, 1945.

Election of Officers, May 29, 1945.

"Graphical Analysis of Frequency Modulation —Part II," by S. H. Van Wambeck, Washington University; June 6, 1945.

SEATTLE

"Kilocycle Cops," by V. A. North, North Pacific Monitoring Area, Federal Communications Commission; June 12, 1945.

WASHINGTON

"Network for Television and Frequency Modulation Broadcasting," by H. S. Osborne, American Telephone and Telegraph Company; February 13, 1945.

"Characteristics and Limitations of the Vacuum-Tube Radio-Frequency Generator," by T. P. Kinn, Westinghouse Electric Corporation; March 12, 1945.

"Wartime Developments in Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; April 9, 1945.

"Radio-Relay Communications System in the United States Army," by W. S. Marks, Signal Corps Ground Signal Agency; May 14, 1945.

"High-Frequency Measurements," by Arnold Peterson, General Radio Company; June 11, 1945.

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"Electrical Glass and Electrical-Glass Working," by E. M. Guyer, Corning Glass Works; June 6, 1945.

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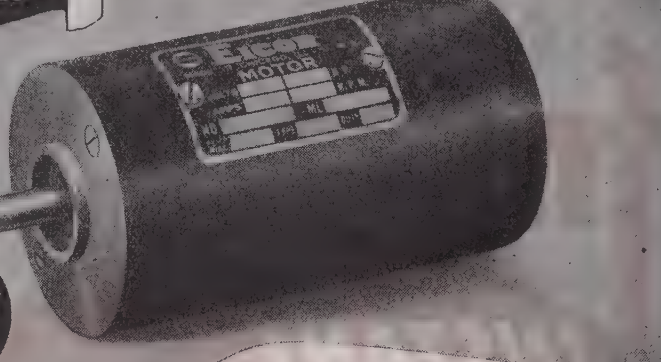

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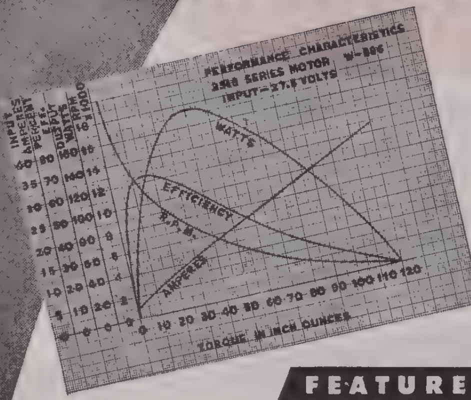

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Lock Torque	(in. oz.)	120	14
Volts Input	(min.)	5	5
Volts Input	(max.)	110	28
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(Continued on page 41A)

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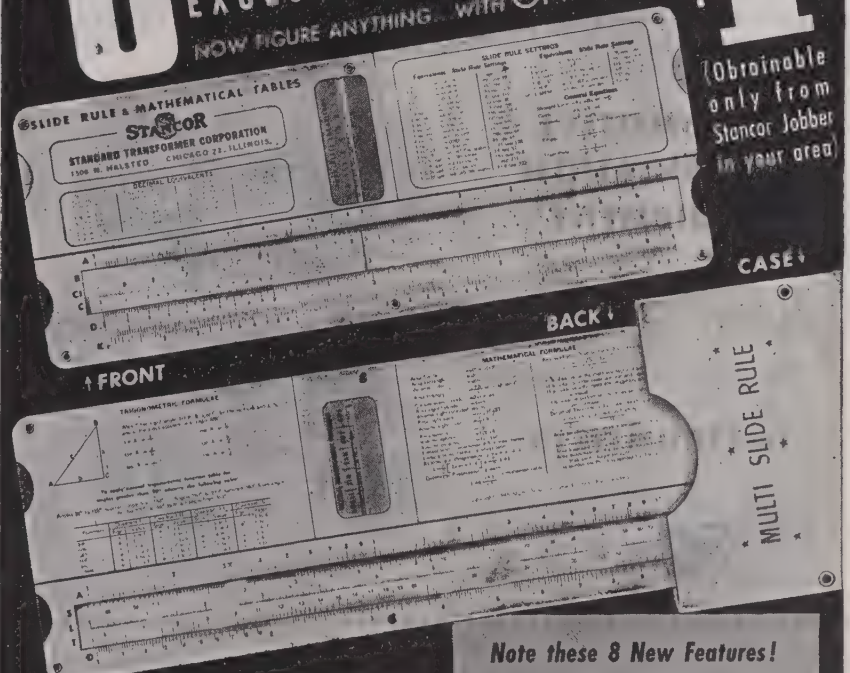

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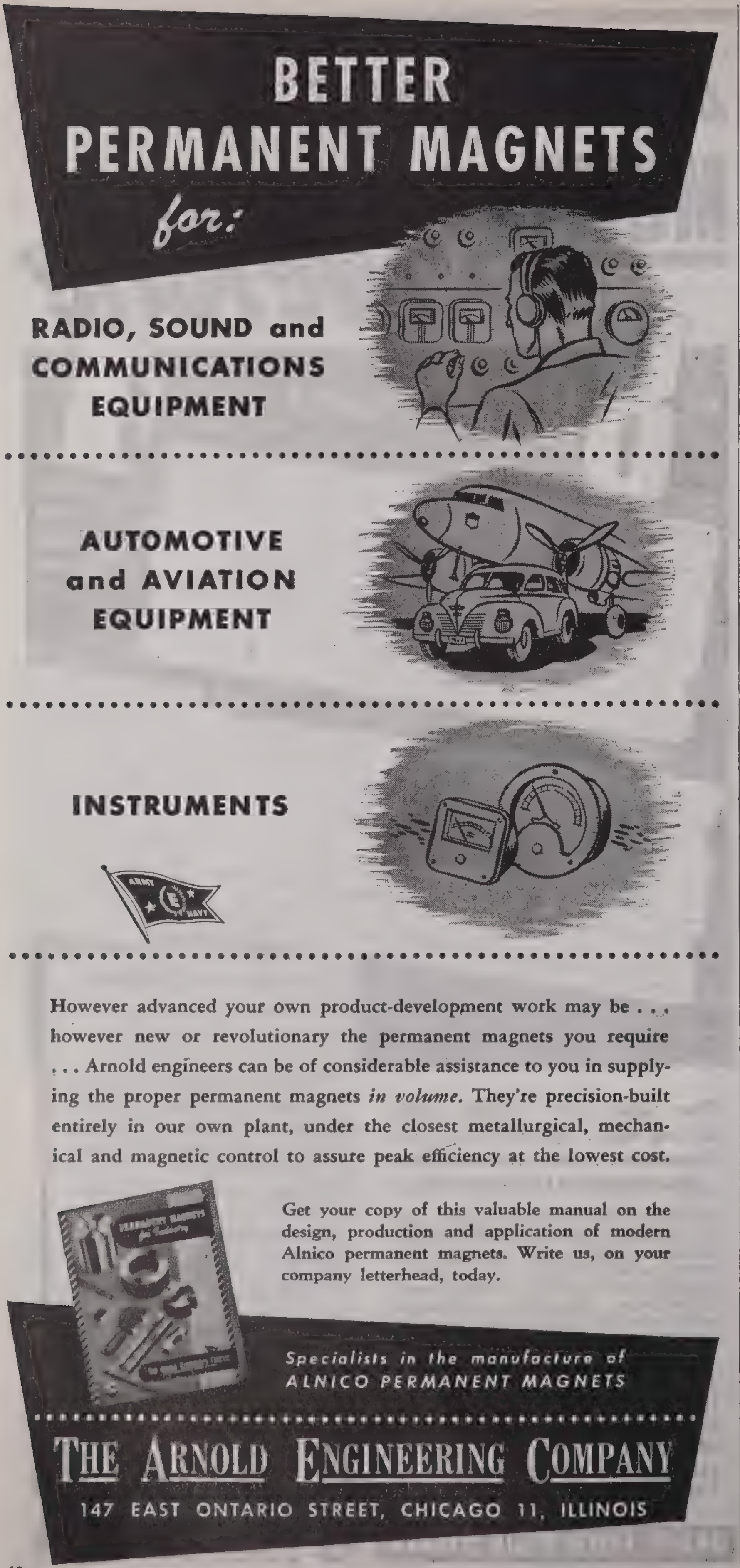

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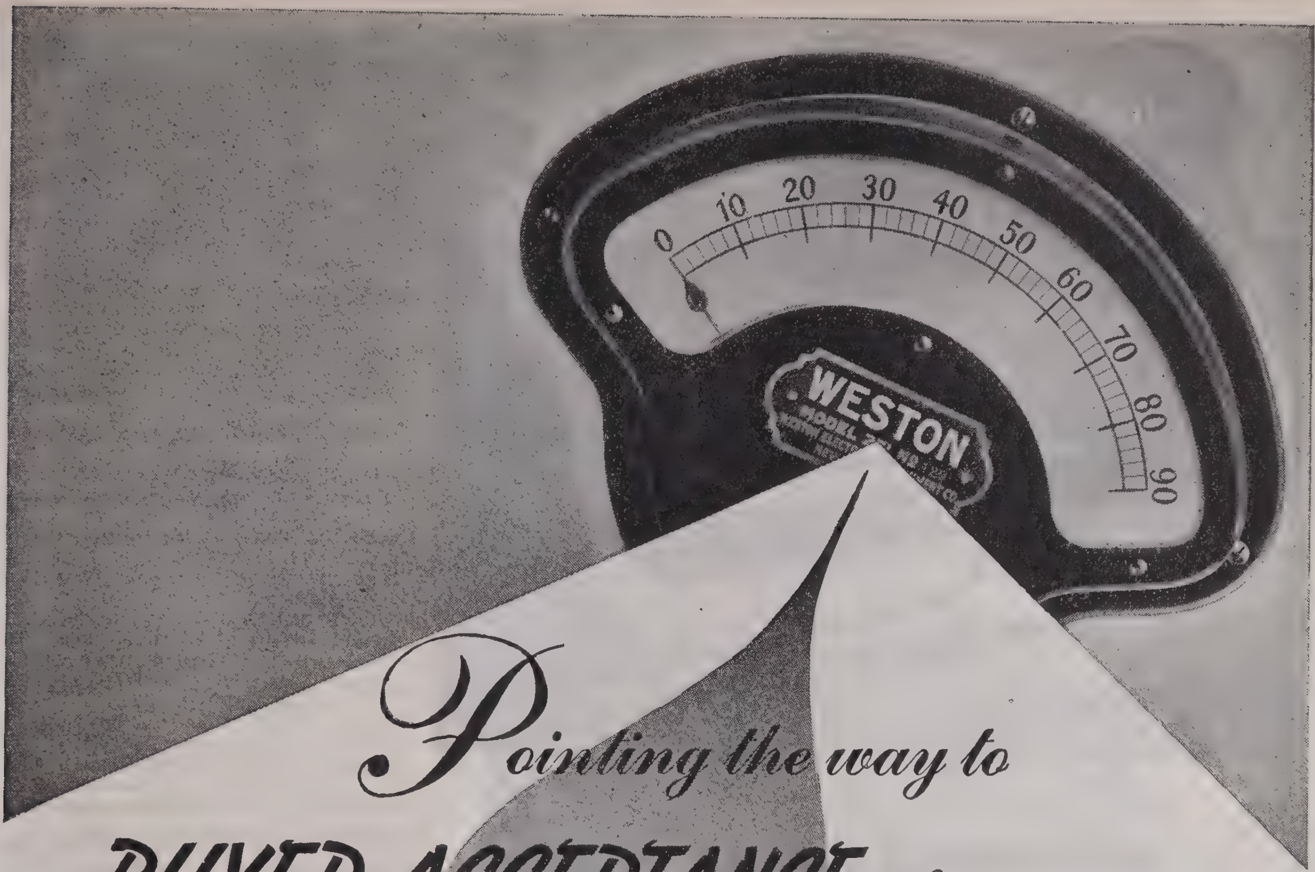
(Continued on page 46A)



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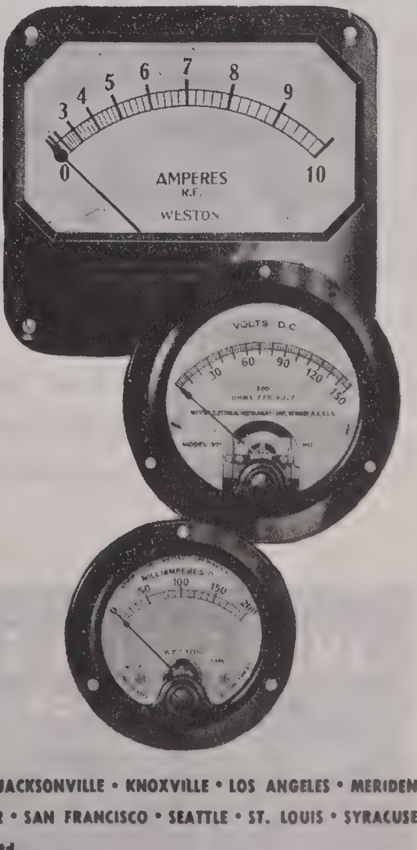

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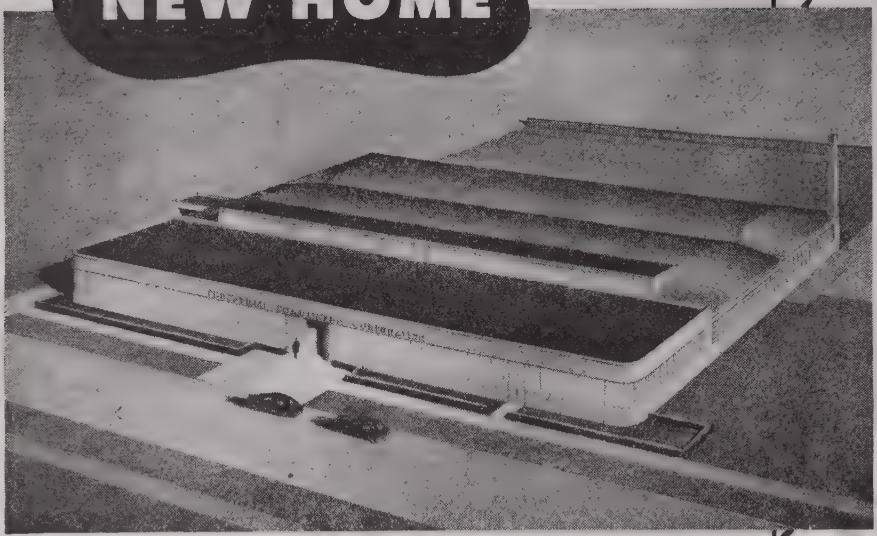

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Rawlins, E. F., 4628 Chester Ave., Philadelphia, Pa.
Reinhart, H. E., RFD 1, Box 442, Meadow Park, San Rafael, Calif.
Rennenkampf, L. J., 30-16—84 St., Jackson Heights, L. I., N. Y.
Rhodes, MacL., 190-06 Hollis Ave., Hollis 7, N. Y.
Rivers, R. A., 1111 Columbus Blvd., Coral Gables 34, Fla.
Rogers, J. S., RFD 2, Box 479-A, Spring Lake, Mich.
Romero, E. S., Fleet Post Office, San Francisco, Calif.
Roselli, D. J., 602—90 St., Brooklyn, N. Y.
Rutherford, R. E., 93 N. Spring Garden, Nutley, N. J.
Sanders, R. G., 900 Riverside Dr., New York 32, N. Y.
Sarli, M. V., General Artigas 4921, Buenos Aires, Argentina
Sarma, P. H. N. B., Wireless Staion, BEZWADA, Madras, South India
Schoew, F.L., II, 1110—10 St., Huntington 1, W.Va.
Schroth, L. C., 50 Colby Rd., North Quincy 71, Mass.
Segaria, N. M., 233 E. Diamond Ave., Hazleton, Pa.
Selke, G. H., 716 Grafton Ave., Dayton 6, Ohio
Shapiro, L., Box 36, Scott Field, Ill.
Silver, B., L.A.A.F., Sqdn Z, Br. 1, Lincoln, Nebr.
Simon, L., 33 Essex St., New York 2, N. Y.
Skrypski, E. J., 73 Belshaw Ave., Eatontown, N. J.
Soich, W. S., 6225 S. Bishop St., Chicago 36, Ill.
Spooner, L. I., 427 Rutherford Blvd., Passaic, N. J.

(Continued on page 48A)

SYLVANIA NEWS
ELECTRONIC EQUIPMENT EDITION

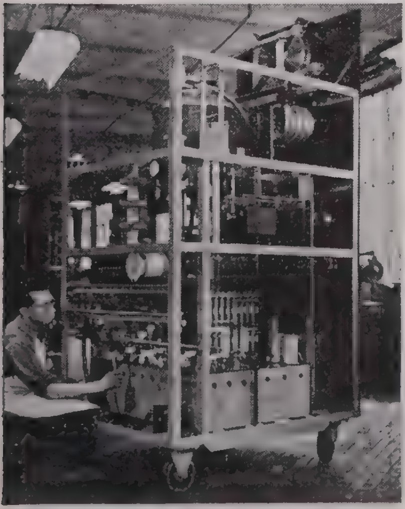
AUGUST *Published by* SYLVANIA ELECTRIC PRODUCTS INC., Emporium, Pa. 1945

HIGH FREQUENCY INDUCTION FURNACE USED IN TUBE PLANT

The bombarder or high frequency induction furnace pictured below is another example of high-precision, modern equipment manufactured at Sylvania Electric's plant in Williamsport, Pa.

Flexible in Application

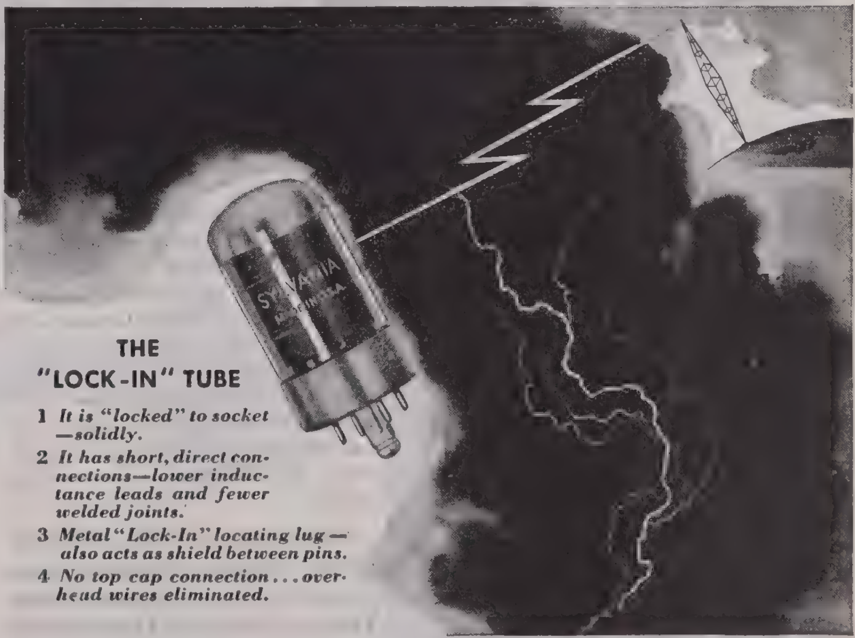
Used in all radio tube plants where exhaust machines operate, this essential apparatus may also be adapted for use in practically any application that requires high frequency induction heating by the connection of the proper heating coils. Its rated input is 25KVA, uses Type 207 tube as oscillator, frequency about 300KC.



High frequency induction furnace used in all radio tube plants where exhaust machines operate. Made by Sylvania Electric at Williamsport, Pa.

LOCK-IN TUBES PERFECTLY IN LINE WITH RECENT FCC DECISION

High Frequency Sets (FM) Will Get Benefit of Tubes' Electrical Superiority



THE "LOCK-IN" TUBE

1. *It is "locked" to socket —solidly.*
2. *It has short, direct connections—lower inductance leads and fewer welded joints.*
3. *Metal "Lock-In" locating lug — also acts as shield between pins.*
4. *No top cap connection . . . overhead wires eliminated.*

Sylvania Electric's revolutionary type of radio tube — the Lock-In — is so mechanically stronger and electrically more efficient that it takes in its stride the recent FCC decision assigning to frequency modulation the band between 88 and 106 megacycles. The basic electrical advantages of the Lock-In construction are ideally suited to the adoption of higher frequencies.

Mechanically it is more rugged because support rods are stronger and thicker—there are fewer welded joints and no soldered joints—the lock-in lug is metal not molded plastic—the elements are prevented from warping and weaving.

Electrically, it is more efficient because the element leads are brought directly down through the low loss glass header to become sturdy socket pins—reducing lead inductance—and interelement capacity.

Today, the many special features of the Sylvania Lock-In Tube are even more up-to-date than when they were introduced in 1938—a fact of increasing importance when considering the numerous postwar developments in the field of communications.

SYLVANIA ELECTRIC

Emporium, Pa.

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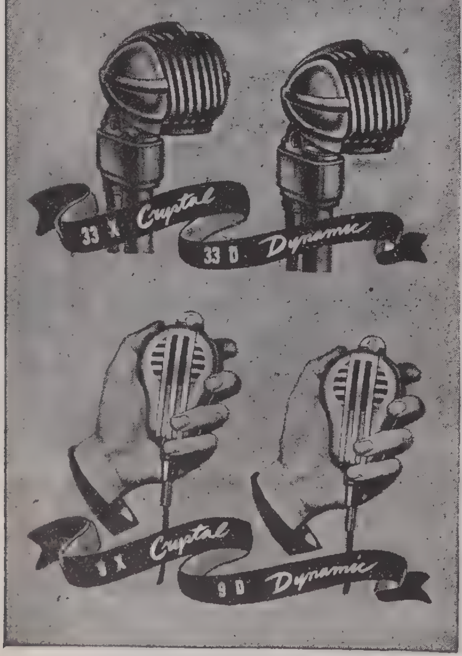
22 D Dynamic



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Turner *Microphones*

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Membership

(Continued from page 46A)

Stubbins, W. F., AAF Training Aids Div., 1 Park Ave., New York 16, N. Y.
Sullivan, F. R., Sylvania Electric Products, Inc., 1221 W. 3 St., Williamsport, Pa.
Taskin, H., 621 N. 26 St., Milwaukee 3, Wisc.
Tong, E., Fleet Post Office, New York, N. Y.
Troise, S., 457—11 Ave., Paterson, N. J.
Updike, A. G., 3425 N. Broad St., Philadelphia, Pa.
Utterback, W. I., Jr., 1810 Kite Ave., Huntington, W. Va.
Van Velsor, C., 174 Lincoln Ave., Little Falls, N. J.
Vencill, G. J., 8421 Colonial La., Clayton 5, Mo.
Volf, C. A., 10 Deshler Ave., Columbus, Ohio
Volz, C., 332 Iris Dr., Pittsburgh 21, Pa.
Ward, O. C., 517 Neilson St., Berkeley 6, Calif.
Wessels, D.H., Press Wireless, Inc., Hicksville, L. I., N. Y.
Wilson, J. H., Jr., 54 W. Blithedale Ave., Mill Valley, Calif.
Wimp, L. F., 7911 Montgall St., Kansas City 5, Mo.
Workman, G. D., 1120 Chestnut St., Kenova, W. Va.
Young, L. L., 1641 K St., N. W., Washington, D. C.
Ziger, A., Bulnes 650, Buenos Aires, Argentina
Zierdt, C. H., Rosedale, Verona, Pa.

ERRATA

The May 2, 1945 Membership Lists, printed in the June issue of the PROCEEDINGS should be corrected as follows:

Page 38A: After J. C. R. Punchard, refer to page 68A beginning with S. D. Robertson, and ending with A. K. Wright for further transfers to Senior Member.

Page 40A: Admissions to Member should end with D. Westwood. Names beginning with R. E. Clinton and ending with C. K. Hadlock should be listed as Associate Members.

Pages 44A, 48A: All names on these pages are Associate Members.

Page 68A: Names beginning with A. W. J. Wessel and ending with R. S. Choate are Associate Members. Names beginning with S. D. Robertson are Transfers to Senior Member as mentioned above.

FRANK MASSA
Electro-Acoustic Consultant
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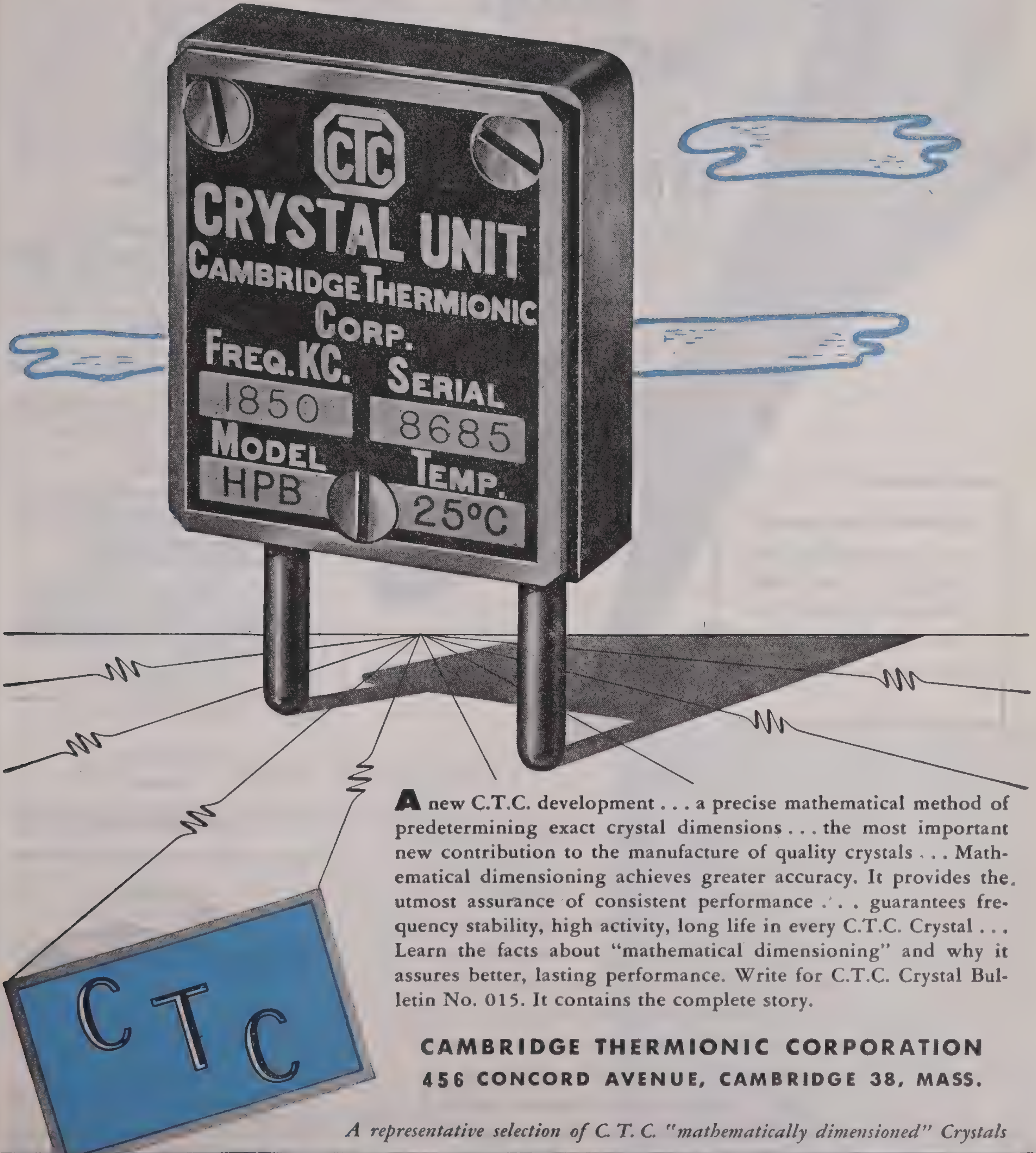
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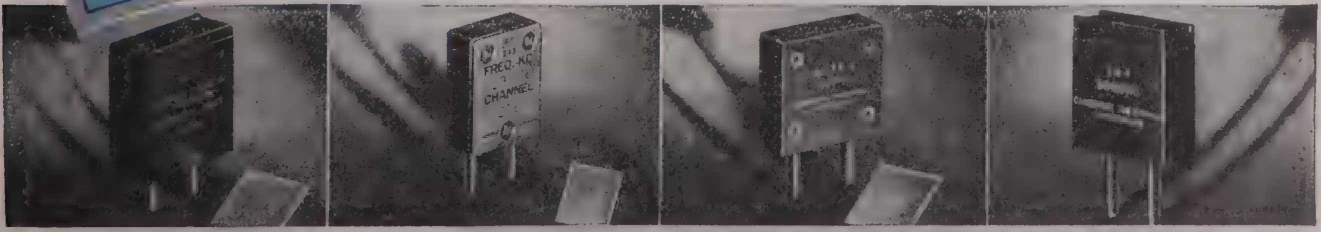
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SECTION MEETINGS

DETROIT

"Wartime Developments of Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; March 14, 1945.

"History of Tube Development," by E. I. Mouromtseff, Westinghouse Electric Corporation; April 19, 1945.

"Land-Mine Detector," by Roger Schell, International Detrola Corporation; May 18, 1945.

"Supersonics—A New Tool for Industrial Research and Control," by J. R. Frederick, University of Michigan; June 15, 1945.

LONDON (Canada)

"Pulse Techniques," by Robert Wilton, RCAF Radio School; June 7, 1945.

Election of Officers; June 7, 1945.

LOS ANGELES

"The Electron Microscope, Basic Principles and Design," by W. V. Houston, California Institute of Technology; June 17, 1945.

Demonstration of the RCA Electron Microscope, by D. C. Pease, University of Southern California, School of Medicine, U.S.C.; July 17, 1945.

PITTSBURGH

"Telephone Service for Vehicles and Vessels in Urban Areas," by R. T. Griffith, Bell Telephone Company of Pennsylvania; May 14, 1945.

Dinner Meeting, June 12, 1945.

Election of Officers, June 12, 1945.

PORTLAND

"The Scanitran System of Highway Sound Tracks," by D. L. Brown, Portland General Electric Company; June 21, 1945.

SAN DIEGO

"Our Electrical Universe," by L. E. Reukema, University of California; July 6, 1945.

SUBSECTIONS

COLUMBUS

"25-Kilowatt-10-Megacycle Oscillator for Cyclotron," by R. B. Jacques; July 13, 1945.

MEMBERSHIP

The following transfers and admissions were approved on August 2, 1945:

Transfer to Senior Member

Belov, F. I., 1610 Park Road, N.W., Washington, D. C.
Cullum, A. E. Jr., Radio Research Laboratory, Harvard University, Cambridge 38, Mass.
Hammond, J. W., 4 Alabama Court, Towson 4, Md.
Heath, F. J., 5 du Maurier Blvd., Toronto 12, Ont., Canada
Holmes, R. S., 2 Ober Road, Princeton, N. J.
Kalbfell, D. C., 941 Rosecrans St., San Diego 6, Calif.
Kolar, E. F., 1622 Shadyside Road, Baltimore 18, Md.
Marshall, C. J., 1 Twain Pl., Dayton 10, Ohio
Monfort, R. A., 7 Birch La., New Hyde Park, N. Y.

(Continued on page 36A)

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We have been at it successfully since 1922—long enough to gain genuine, practical know-how in every phase of the business. That experience, plus engineering ability and precision workmanship, add up to the kind of Equipment which appeals to Engineers and Station Managers alike.

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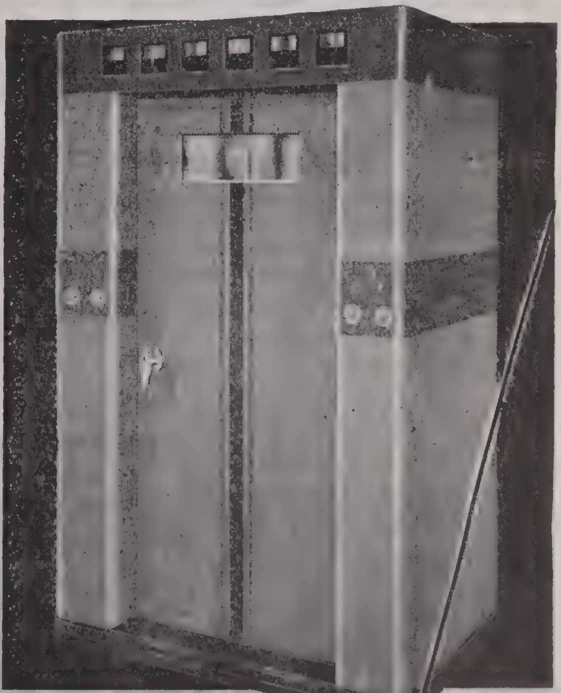

PROGRESS REPORT

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EXCLUSIVE MANUFACTURERS OF RADIO TRANSMITTING EQUIPMENT SINCE 1922

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Condensed Power for Years of Service

VERSATILITY and dependability were paramount when *Alliance* designed these efficient motors — *Multum in Parvo!* . . . They are ideal for operating fans, movie projectors, light home appliances, toys, switches, motion displays, control systems and many other applications . . . providing economical condensed power for years of service.

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Our long established standards of precision manufacturing from highest grade materials are strictly adhered to in these models to insure long life without breakdowns.

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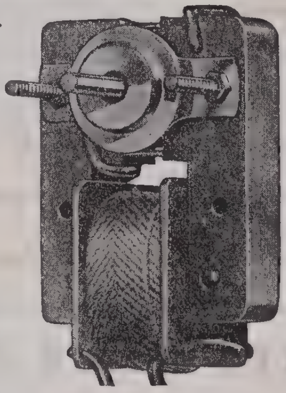
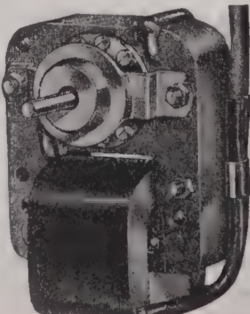


Model "MS"—Full Size Motor Measures 1¾" x 2 x 3⅛"



New Model "K"—Full Size Motor Measures 2⅛" x 2⅜" x 3 3/16"

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—YOUR ALLY IN WAR AS IN PEACE

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MANUFACTURING CO.
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MEMBERSHIP

(Continued from page 34A)

Paine, R. C., 436 Cornelia St., Boonton, N. J.
Smith, L. C., 337 Westmont Ave., Westmont, Collingswood, N. J.
Talley, D., 130 Martense St., Brooklyn, N. Y.
Tellier, J. C., 233 Crosshill Road, Philadelphia 31, Pa.
Warren, J. R., 21 Otter Crescent, Toronto 12, Ont., Canada
Wilson, E. S., 1265 Culver Rd., Rochester 9, N. Y.

Admission to Senior Member

Hicks, C. W., Phoenix, Md.
Stoner, F. E., 2737 Devonshire Pl., N.W., Washington, D. C.

Transfer to Member

Antman, M. A., General Delivery, Dayton 1, Ohio
Babcock, W. C., 213-05—34 Road, Bayside, L. I., N. Y.
Bertolet, E. C., 5135 Marion St., Germantown, Pa.
Bird, L. T., 318 Monmouth Ave., Montreal, Que., Canada
Blaisdell, H. L., 52-22 Marathon Pkwy., Little Neck, L. I., N. Y.
Dawes, H. H., General Radio Co., 275 Massachusetts Ave., Cambridge, Mass.
Evans, C. B., 1435 Snowden Ave., Memphis 7, Tenn.
Friedman, E. D., Federal Telephone and Radio Corporation, 200 Mount Pleasant Ave., Newark, N. J.
Gerken, G. H., 11701 Bullis Road, Lynwood, Calif.
Grace, J. R., 1323 Second Ave., San Mateo, Calif.
Hartwell, H. G., 1405 Stewart Ave., Huntington, West Va.
Hatfield, L. N., 4405 First Pl., N.E., Washington 11, D. C.
Hershfield, S., 22 State Normal Pl., Jersey City 5, N. J.
Labby, D. M., 3157 Maple St., San Diego 2, Calif.
Lundahl, T., 516 W. 34 St., New York 1, N. Y.
Piersall, B. P., U. S. Naval Ordnance Plant, Forest Park, Ill.
Powers, A. B., Box 44, Kent, Wash.
Randall, E. F., 1947 Broadway, Studio 532, New York, N. Y.
Rensch, H. A., 7649 Eastlake Ter., Chicago 26, Ill.
Renton, R. J., 1416 N. Longfellow St., Arlington, Va.
Robinson, E. F. V., Radio Branch, National Research Council, Ottawa, Ont., Canada
Schmitz, W., 451 Main St., Waltham, Mass.
Stenger, F. J., 7828 Truxton Ave., Los Angeles 43, Calif.
Stewart, A. W., 19508 Greenfield Rd., Detroit 19, Mich.
Tiley, J. W., 4226 Comly St., Philadelphia 24, Pa.
Weber, C. E., 8364 Jadwin St., Cincinnati 15, Ohio
Williams, L. A., RFD 2, Box 43, Anchorage, Ky.
Wilson, W. R., 5 Enjay Ave., Baltimore 28, Md.

Admission to Member

Baillie, W. L., British Admiralty Delegation (D.R.E.-W.) Box 165, Benjamin Franklin Station, Washington, D. C.
Boyd, B., 4733 Rocky River Dr., Cleveland 11, Ohio
Coffey, L. E., 36 Nepean St., Ottawa, Ont., Canada
Curtis, H. W., Dept. of Chemistry and Electricity, U.S.M.A., West Point, N. Y.
Hawkins, W. G., 98 Rose St., Freeport, L. I., N. Y.
Kennedy, D. D. Jr., c/o Postmaster, New York, N. Y.
Price, J. A., 36 Simeon St., Kitchener, Ont., Canada
Selgin, P. J., 10 Albert Pl., New Rochelle, N. Y.

(Continued on page 38A)

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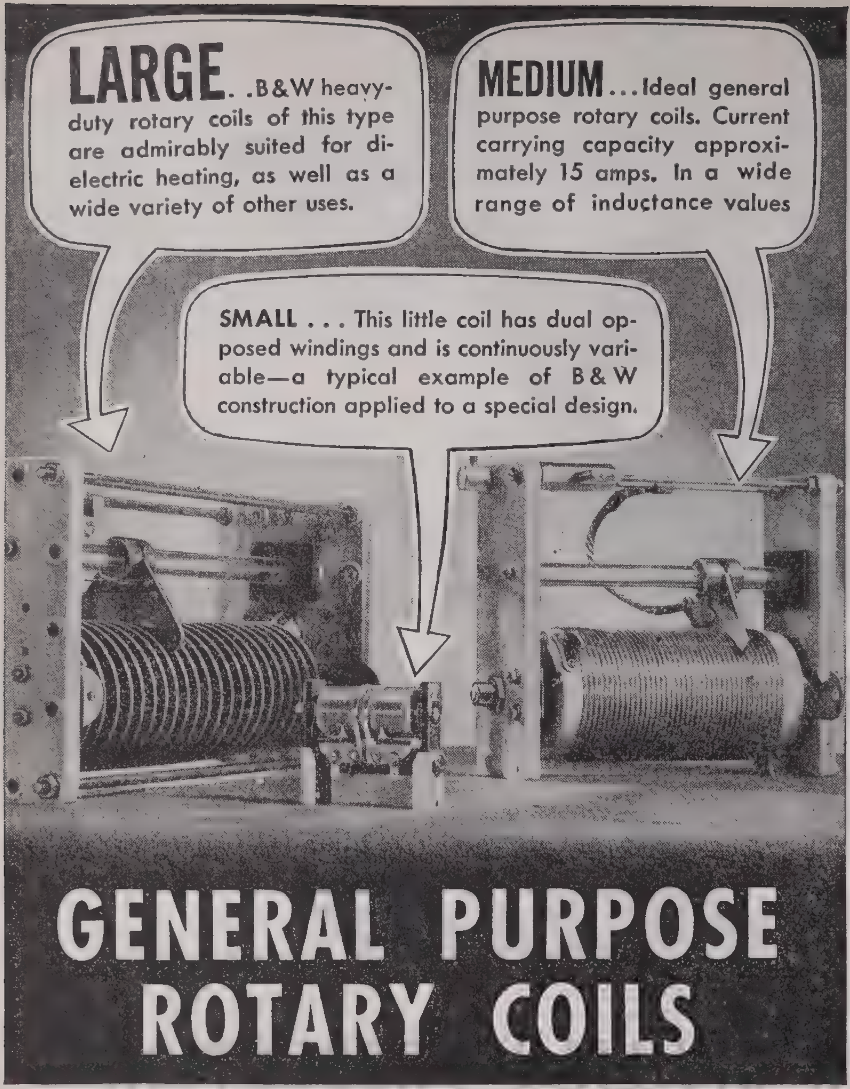
Factories in Dobbs Ferry, N.Y.; Mount Vernon, N.Y. (Metalix Div.); Lewiston, Me. (Elmet Div.)



LARGE. . B&W heavy-duty rotary coils of this type are admirably suited for dielectric heating, as well as a wide variety of other uses.

MEDIUM. . . Ideal general purpose rotary coils. Current carrying capacity approximately 15 amps. In a wide range of inductance values

SMALL . . . This little coil has dual opposed windings and is continuously variable—a typical example of B & W construction applied to a special design.

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ROTARY COILS

. . . in a wide range of inductances, voltages and capacities

Typical of the completeness and engineering supremacy of the general line of B&W Air Inductors, these rotary coils are available in standard or special types, sizes, and shapes to meet practically any rotary coil requirement. Bring your coil problems to coil headquarters! B&W engineers will gladly make recommendations based on an unparalleled experience in this specialized, rapidly advancing field.

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MEMBERSHIP

(Continued from page 36A)

Sprinkle, R. E., 974 School St., Box 461, Webster, Mass.
Wood, D. F., 201 Miami Club Dr., Mishawaka, Ind.

Admission to Associate

Anderson, A. G., c/o Fleet Post Office, San Francisco, Calif.
Asendorf, J. A., 166 N.E. 28 St., Miami 37, Fla.
Barrows, J. H. Jr., 6870 N. Sedro St., Portland 3, Ore.
Bartlett, F. E., KSO, Des Moines 9, Iowa
Bayer, S., c/o Fleet Post Office, San Francisco, Calif.
Beardsley, W. S., 8648 Georgia Ave., Silver Spring, Md.
Becker, L. S., 385 S. Burnside Ave., Los Angeles 36, Calif.
Belgamvala, M. K., 15-A Wheatcroft Road, Nungambakkam, Madras, India
Bell, H. C., 314 Lee Ave., Toronto, Ont., Canada
Bencivenga, P., Illustrated Technical Products, 2 Lafayette St., New York, N. Y.
Billman, E. F., 95 Sherwood Ave., Teaneck, N. J.
Blank, P., 789—28 Ave., San Francisco, Calif.
Blatt, J. M., RCA Laboratories, Princeton, N. J.
Bradley, J. L., Newell-Emmett Co., 40 E. 34 St., New York 16, N. Y.
Buckwitz, S., 15081 Littlefield St., Detroit 27, Mich.
Bullington, K., 463 West St., New York 14, N. Y.
Bundrick, R. J., 1622 Merced St., Richmond, Calif.
Butler, R. E. Jr., 925 Jones St., San Francisco 9, Calif.
Cairns, J. R., 348 Lumsden Ave., Toronto 13, Ont., Canada
Cannizzaro, M., Radio Materiél School, Staff, Treasure Island, San Francisco, Calif.
Carnes, E. C., 215 S. Walnut St., Springfield, Ill.
Chace, W. G., 205 Ashland Ave., West Manayunk, Pa.
Clark, W. E., Sonar Maintenance Staff, West Coast Sound School, San Diego 47, Calif.
Cohoon, B. C., Radio Materiél School, Treasure Island, Calif.
Colburn, O. E., 3119 Chestnut St., Long Beach, Calif.
Creamer, E. M. Jr., 1313 Medary Ave., Philadelphia 41, Pa.
Dalrymple, I. N., Staff-Radio Materiél School, Treasure Island, Calif.
Dare, E. B., 1723 E. 4 St., Mishawaka, Ind.
Dayton, D. M., 2525 Durant Ave., Berkeley 4, Calif.
Delmege, A. H., 2310 High St., S.E., Washington 20, D. C.
Dewey, O. L., Box 236, Concord, Calif.
Diaz, E. R., Bendix Radio, Townson, Md.
Doba, S. Jr., 180 Varick St., New York, N. Y.
Dodd, G. M., 1114 N. Stafford St., Portland 3, Ore.
Dopkowski, W. J., Sonar Maintenance Staff, West Coast Sound School, San Diego 47, Calif.
Duncan, J. A., 4041—39 Ave., S.W., Seattle 6, Wash.
Elsken, R. H., 925 Chenery St., San Francisco, Calif.
Emm, T. T., 1552 Pershing Ave., San Bernardino, Calif.
Friend, J. L. Jr., 73 Tanager St., Arlington 74, Mass.
Gilleo, W. L., 318 N. Broad St., Peekskill, N. Y.
Genovese, A. N., 3917 Fullerton Ave., Chicago 47, Ill.
Goldberg, H., 1219 Union St., Brooklyn 25, N. Y.
Gorman, W. F., 1198 E. 94 St., Brooklyn 12, N. Y.
Greenwald, I. H., 4526 N. Hermitage St., Chicago 40, Ill.
Guichon, L., Northern Electric Co., Box 369, Montreal, Que., Canada

(Continued on page 40A)

For your problems of induction heating . . .

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CAPACITORS give you the precise capacitance value you want

Wherever *exact* capacitance values are needed . . . where exceedingly close tolerances are demanded . . . choose I.C.E. Capacitors. For induction heating . . . machine speed controllers . . . for current rectifiers . . . for any industrial application . . . I.C.E. Capacitors provide greater operating safety.

I. C. E. PRECISION GRADE VACUUM CAPACITORS

Value Range	*Accurate to*
6 mmfd. to 25 mmfd.	±0.5 mmfd.
26 mmfd. to 60 mmfd.	±1.0 mmfd.
61 mmfd. to 110 mmfd.	±1.5 mmfd.

I. C. E. XX GRADE VACUUM CAPACITORS

Value Range	*Accurate to*
6 mmfd. to 25 mmfd.	±0.2 mmfd.
26 mmfd. to 60 mmfd.	±0.3 mmfd.
61 mmfd. to 110 mmfd.	±0.5 mmfd.

Close Tolerances With I.C.E. Capacitors

Besides offering you a wide range of capacitance values, I.C.E. vacuum capacitors are built to give you previously unobtainable tolerances for extreme precision control

Full Details in the New I. C. E. Catalog

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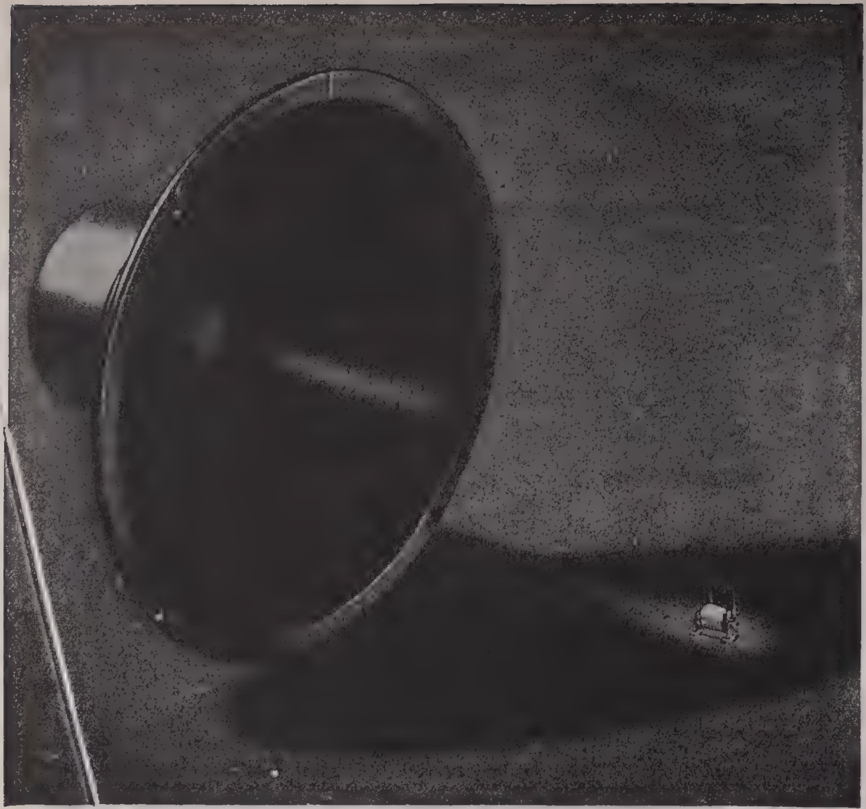

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Hallmark, G. D., Box 4798 College Station, Texas
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Hart, R. N., 7322 Bryan St., Philadelphia 19, Pa.
Hartley, J. M., RMS Staff, Treasure Island, San Francisco, Calif.
Henry, J. C., 186 Jersey St., San Francisco 14, Calif.
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Hoyle, W. G., c/o D.E.E., N.S.H.Q., Ottawa, Ont., Canada
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Ingerson, K. E., 9 Euclid Circ., E. Natick, Mass.
Jackson, W. H., RMS Staff-10, Treasure Island, San Francisco. Calif.
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Kettinger, L. H., 3376 Grove St., Berkeley 3, Calif.
Kinard, J. M. Jr., 825—53 St., Oakland, Calif.
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Kirk, H. E., 1315—25 St., Bakersfield, Calif.
Klopp, A. E., 5705 London Rd., Duluth 4, Minn.
Knauss, H. C., 30 Lancaster St., Cambridge 40, Mass.
Knipe, W. M., RMS Staff, Treasure Island, Calif.
Kochenderfer, W. E. Jr., 3443 Scott St., San Francisco 23, Calif.
Korin, S. B., 1735 Fulton Ave., Bronx 57, N. Y.
Lackey, B. B., 1832 Masonic Ave., San Francisco, Calif.
Langer, W. A., Box 293, Alexandria, Minn.
Leach, P. C., Sonar Maintenance Staff, West Coast Sound School, San Diego 47, Calif.
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Lewis, G., 1490 Jesup Ave., New York 52, N. Y.
Lucas, H. L., 2099 Willamont Ave., Columbus, Ohio
Lynch, H. S., 336 Beacon St., Boston, Mass.
Malta, S. V., 8705 Raven Dr., Towson 4, Baltimore, Md.
Marnitz, J. W., S.S.S., S.A.C.S., Box 4919, Johannesburg, South Africa
McClung, R. R., 1012 N. Kelham Ave., Oklahoma City 4, Okla.
McLaughlin, W. F., RMS Staff, Treasure Island, San Francisco, Calif.
Meadows, E. A., 430 S. Clark Dr., Beverly Hills, Calif.
Mee, R. W., 1802 Blake St., Berkeley 3, Calif.
Miller, A. T., 1357 W. First Ave., Columbus 8, Ohio
Moore, C. N., Radio Materiél School Staff, Treasure Island, Calif.
Moore, W. M., 306 E. Market St., Sandusky, Ohio
Morse, S. P., 425 W. 23 St., New York 11, N. Y.
Moulic, R. J., 1003 E. Jefferson St., Bloomington, Ill.
Munster, A. C., 264 W. Sulis St., Philadelphia 20, Pa.

(Continued on page 42A)

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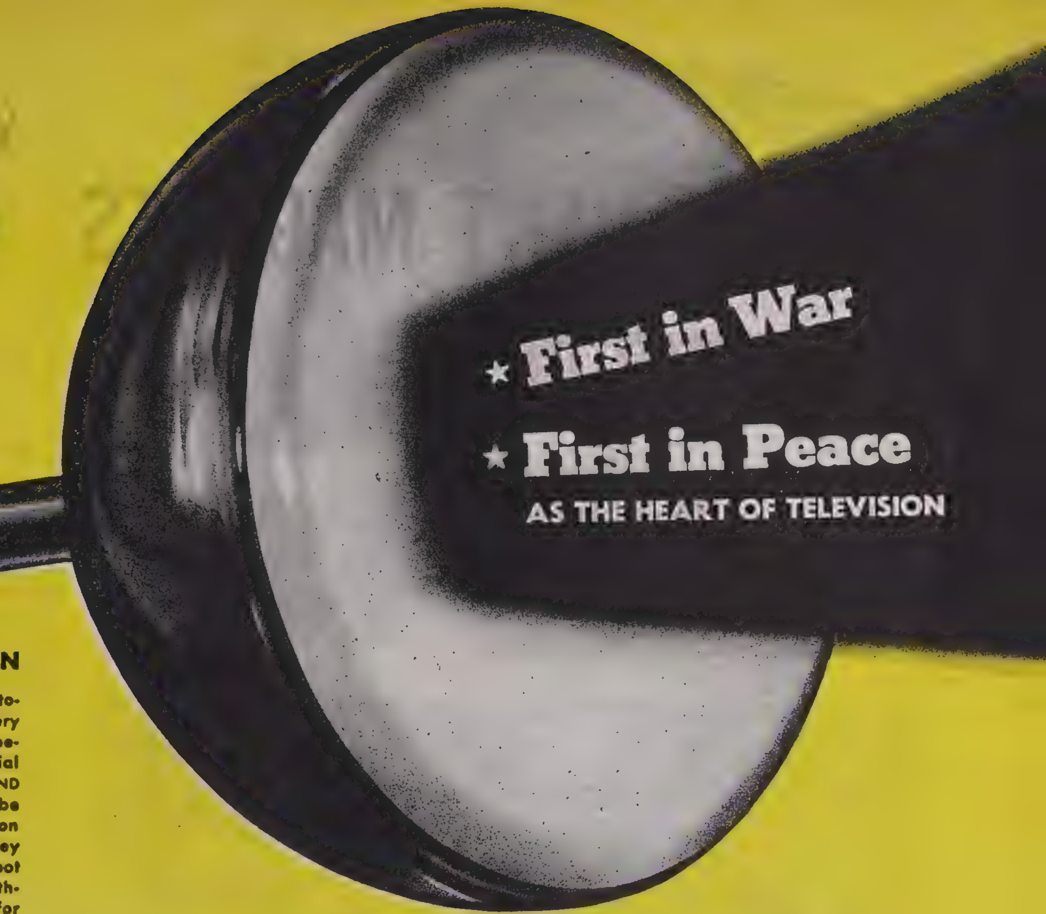

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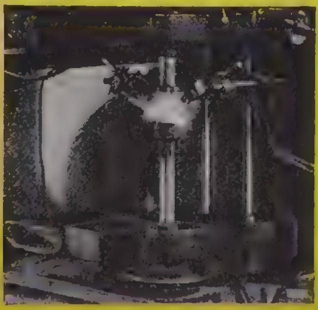
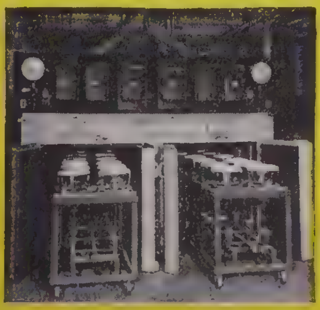

Manufacturing header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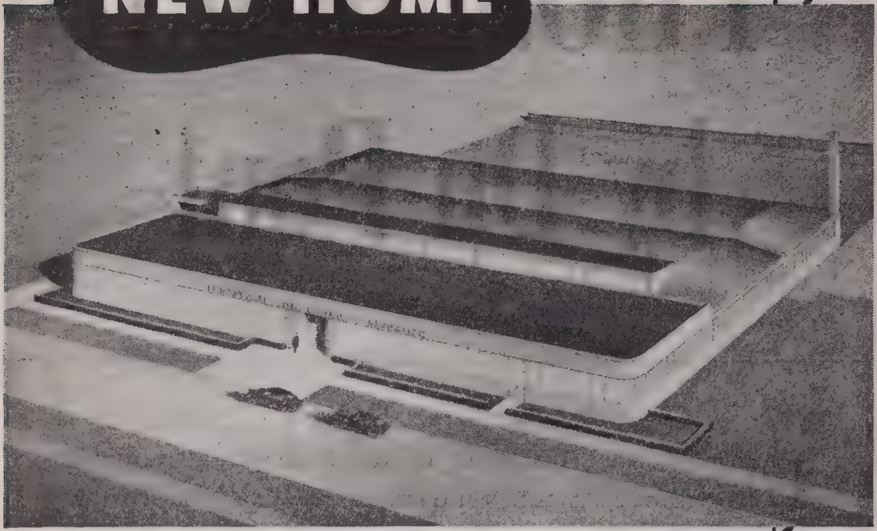

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(Continued from page 42A)

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Swingle, D. M., 612 Seventh Ave., Asbury Park, N. J.
Sze, F. C., International Training Administration, 734—15 St., Washington, D. C.
Talley, J. P. Jr., KRLD Transmitter, Garland, Texas
Teeple, A. L., 2327 W. Medill Ave., Chicago 47, Ill.
Thomas, G. G., Bell Telephone Laboratories, Murray Hill, N. J.
Thompson, M. I., Box 1284, Oak Ridge, Tenn.
Thompson, R. L., 2091 California St., Berkeley 3, Calif.
Thuis, D. L., 1733 Bancroft Way, Berkeley, Calif.
Truszynski, G. M., 155 Shenandoah Rd., Hampton, Va.
Updegraff, E. E., WMT Transmitter, RFD 3, Marion, Iowa
Vendely, B., 220 W. Herman St., Dayton 5, Ohio
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Vitzelio, R. B., RMS Staff, Treasure Island, San Francisco, Calif.
Vollintine, K. R., 6400 Seventh Ave., Los Angeles, Calif.
Waddell, C. N., Instruction Staff RMS, Treasure Island, Calif.
Wakefield, S. J., 600 S. Lyman St., Oak Park, Ill.
Walker, L., Hotel Philadelphian, 39 at Chestnut St., Philadelphia, Pa.
Washburn, E. L., Electronics School, Camp Wawona, Wawona, Calif.
Watlington, F. W., 1 Rose Blvd., Baldwin, N. Y.
Wells, T. H., Box 771, 7166 Fay Ave., La Jolla, Calif.
Weyand, C. L., New Haven Rd., Prospect, RFD 2, Waterbury, Conn.
Whitman, M. R., 1959 Lunt St., Chicago 26, Ill.
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Woodley, E. H., 1261 Shearer St., Montreal, Que., Canada

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. . . for dielectric heating applications are available in 25 different types, operating with remarkable efficiencies at frequencies ranging from 20 to 120 megacycles. Write for the Amperex catalog.

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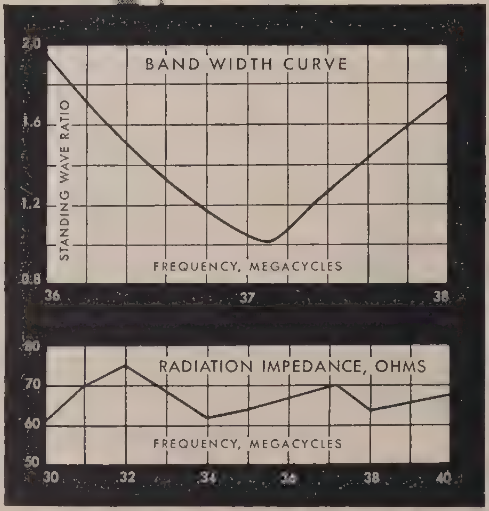

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- Light weight — only 15 pounds — simplifies installation.
- Lightning hazard minimized by grounded vertical element.
- "Slide trombone" calibration permits exact adjustment for any frequency between 30 and 40 MC, using only a wrench. Optimum performance for that frequency is guaranteed without "cut and try" methods.
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SECTION MEETINGS

BOSTON

"Wartime Developments in Electronics," by W. L. Everitt, President, The Institute of Radio Engineers, Inc.; May 25, 1945. Election of Officers, May 25, 1945.

BUENOS AIRES

"Frequency-Modulation Transmitters," by A. Saenz, RCA Victor Argentina; June 1, 1945.

"Television," by J. P. Calvelo; Tel-Rad, S.A.; June 15, 1945.

"Acoustic Proof-Room—General Theory—Loudspeakers Tests' Instrumental," by R. Hastings; Ucoa, S. A.; July 6, 1945.

"Some Points of View on the International Radio-Conferences," by A. T. Cosentino; Pan-American Argentine; July 20, 1945.

CEDAR RAPIDS

"Frequency Ranges in Music and Speech," by Ralph Glover; Consulting Engineer; June 22, 1945.

DALLAS-FT. WORTH

"An Informal Discussion of the RCA System of Television," by Wesley Turner, RCA Victor Division; July 13, 1945.

"Television Equipment Demonstration," by J. C. Skinner, Interstate Circuit, Inc.; July 13, 1945.

PORTLAND

"Frequency Modulation," by those present; August 8, 1945.

SAN DIEGO

"A 500-Cycle Aircraft-Landing System," by E. N. Dingley, U. S. Navy Radio and Sound Laboratory; August 21, 1945.

Election of Officers, August 21, 1945.

TORONTO

"Application of Radio to Railroad Signaling Systems," by W. S. Halstead, Halstead Traffic Communication Corp.; May 21, 1945.

Election of Officers, May 21, 1945.

WILLIAMSPORT

"Radio Relay Systems Development by the Radio Corporation of America," by H. O. Peterson, RCA Laboratories, Inc.; July 27, 1945.

"Industrial Radio-Frequency Transformers," by H. W. Parker, Sylvania Electric Products, Inc.; July 27, 1945.

"The Use of Sound Systems as a Production Tool," by A. G. Schifino, Stromberg-Carlson Company; July 28, 1945.

MEMBERSHIP

The following transfers and admissions were approved on September 5, 1945:

Transfer to Senior Member

Bearden, J. A., Johns Hopkins University, Radiation Laboratory, Baltimore 18, Md.
Bloomer, T. M., 605 Warwick Road, Baltimore 29, Md.
Bonney, R. B., 343 Knollwood Ave., Douglaston L.I., N.Y.
Dickinson, W. A., Sylvania Electric Products, In Emporium, Pa.
Downs, G. W., California Institute of Technolo 137 South Lake Ave., Pasadena, Calif.
Ingalls, C. E., 87 Simonds Road, Lexington Mass.
Johnson, W. C., School of Engineering, Princ N.J.

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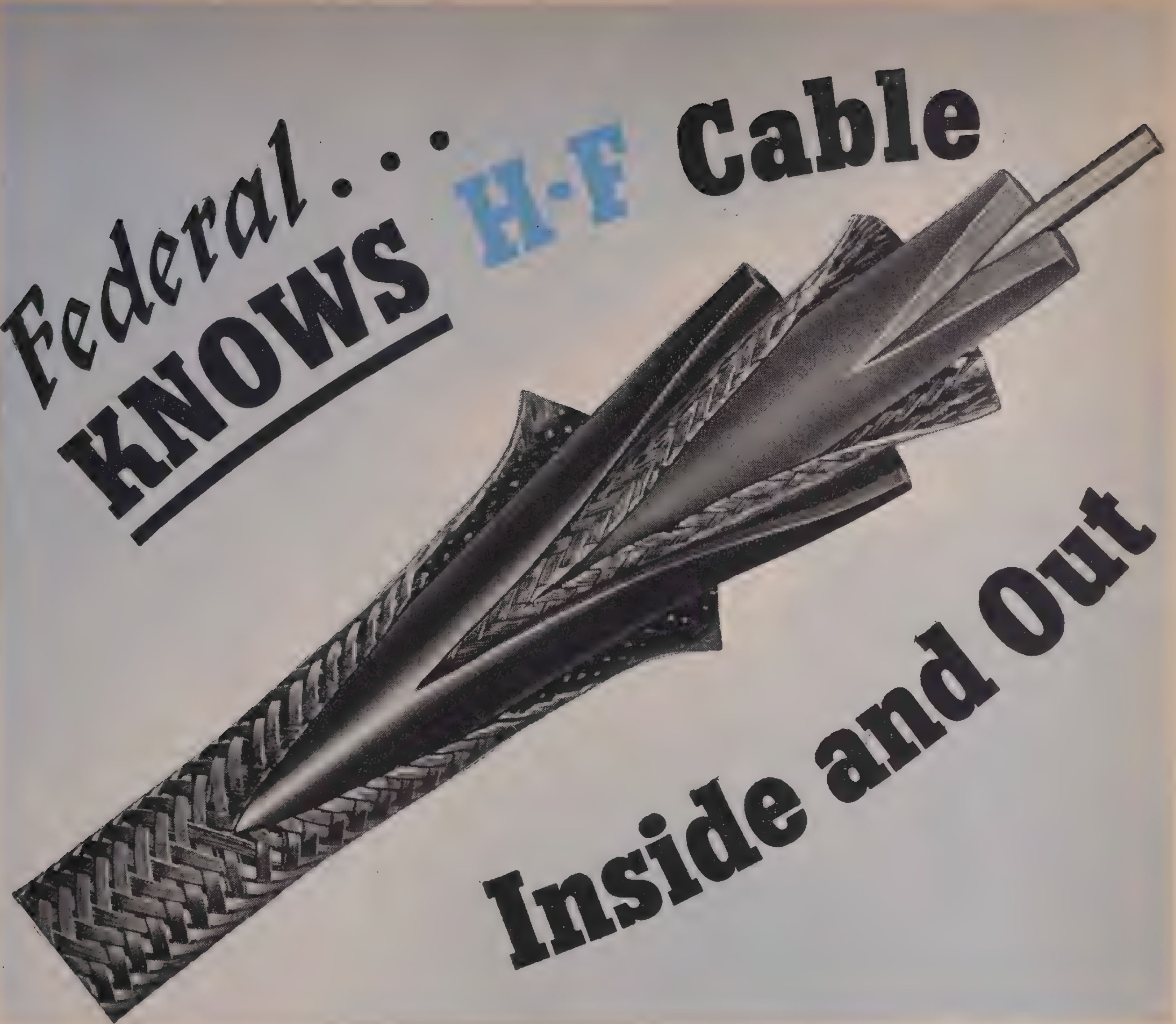

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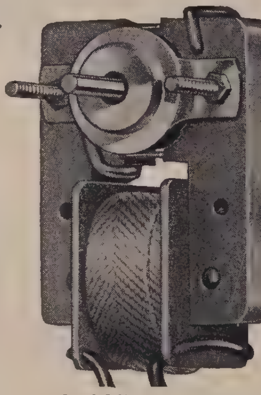
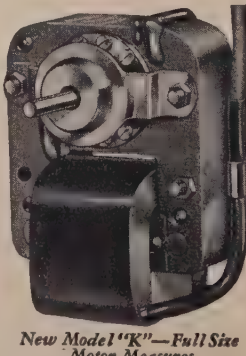


Model "MS" — Full Size Motor Measures 1¾" x 2 x 3⅛"



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Lewis, O. I., 128 E. Custis Ave., Alexandria, Va.
McCanne, L., Stromberg-Carlson Co., Rochester 3, N.Y.
Mueller, C. W., 17 Harold Ave., Princeton, N.J.
Muller, J. H., 23 Johnstone Rd., Great Neck, L.I., N.Y.
O'Brien, E. J., 960 Mamaroneck Ave., Mamaroneck N.Y.
Phelan, T. H., 364 Concord Drive, Maywood, N.J.
Ready, W. A., 85 Chestnut St., Weston, Mass.
Young, C. J., RCA Laboratories, Princeton, N.J.

Admission to Senior Member

Ferrar, R. C., P. O. Box 315, New Providence, N. J.
Grignon, L. D., 1427 Warnall Ave., Los Angeles 24, Calif.
Walker, H., 506 Overbrook Rd., Baltimore 12, Md.

Transfer to Member

Anderson, I. H., 1709—3 Ave., S., Anoka, Minn.
Baily, F. A. A., 2850 Willowdale Ave., Montreal, P.Q., Canada
Bailey, F. M., General Electric Co., Bldg. 269, Schenectady, N. Y.
Bartman, F. L., 174 N. Arlington Ave., East Orange, N. J.
Berg, R. A., 138 Lexington Ave., Fair Haven, N. J.
Brown, R. H., 1418 M St., N.W., Washington 5, D. C.
Bullock, M. W., 1100 Idylwild Drive, Lincoln 3, Nebr.
Evans, H. R., 1130 N. 9 St., Lafayette, Ind.
Feindel, A., 292 N. Maple Ave., East Orange, N. J.
Foster, J. D., 30 Burns Ave., Wyoming 15, Ohio
Gaenzle, J. B., 301 E. Allegany Ave., Emporium, Pa.
Goldman, A., 565 W. 2 St., Dayton 2, Ohio
Holling, K., "Lynwood," Priestsic Rd., Sutton in Ashfield, Notts., England
Jensen, W., 1369 W. Iowa Ave., St. Paul 4, Minn.
Jordan, A. D., 605 Jordan Ave., Montoursville, Pa.
Kingsley, R. J., 180 Sanders Rd., Buffalo, N. Y.
Knowles, D. J., RFD 2, Emporium, Pa.
Lambert, R. L., Sylvania Electric Products, Inc., Emporium, Pa.
Loeber, C. W., Federal Communications Commission, Washington 25, D. C.
Mason, J. E., RFD 3, Box 1121, Bremerton, Wash.
McKenzie, A. A., 350 Cabot St., Newtonville 60, Mass.
McNaughten, N., International Div., Federal Communications Commission, Washington 25, D. C.
Mendenhall, G. G., 1602 E. Garfield St., Seattle 2, Wash.
Morley, H. B., 76 Forster St., Brighton 35, Mass.
O'Brien, R. S., 12 Hilliard St., Cambridge 38, Mass.
Peterson, A. W., 519 E. Allegany Ave., Emporium, Pa.
Porsch, A. W., Box 87, Emporium, Pa.
Raymond, G. H., Upsal Manor, 13 E. Upsal St., Germantown, Philadelphia 19, Pa.
Reitz, N. J., Sylvania Electric Products, Inc., Emporium, Pa.
Roberts, S., 10 Avon St., Cambridge 38, Mass.
Schnering, C. J., 1122—11 Ave., Huntington, W. Va.
Sherburne, C. J., 3139 Wynwood Lane, Los Angeles 23, Calif.
Soderman, R. A., c/o Radio Research Laboratory, 18 Divinity Ave., Cambridge 38, Mass.
Switzenberg, E. A. F., c/o Western Electric Co., D-9813, 195 Broadway, New York 7, N. Y.
Teommey, G. H., Jr., 88 Hart St., Lynbrook, N. Y.
Wheeler, B. F., 500 Redman Ave., Haddonfield, N. J.
Woodbury, W. F., 8 Alden Ave., Valley Stream, N. Y.

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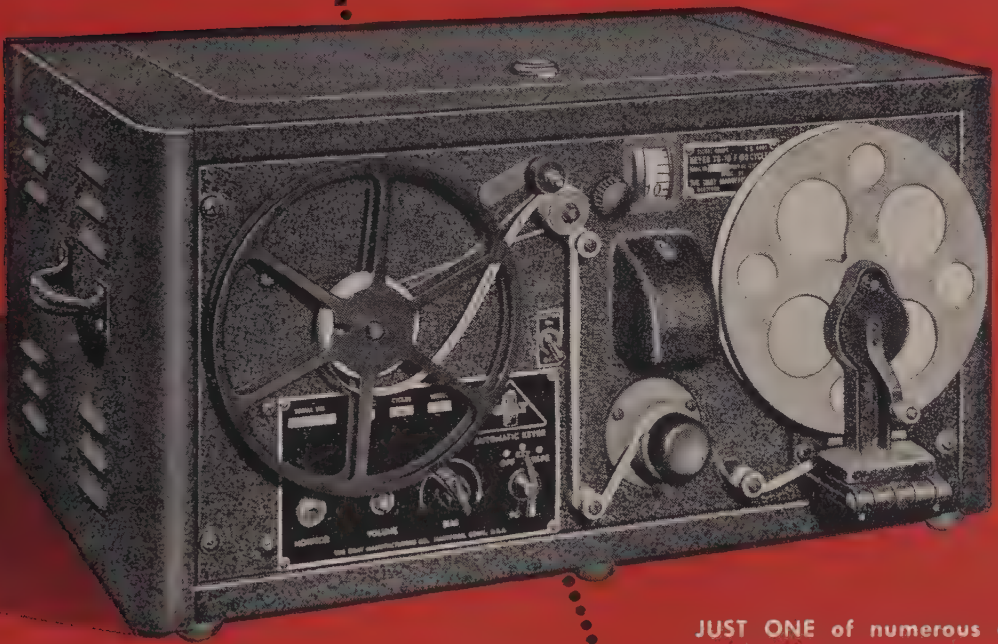
JUST ONE



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MEMBERSHIP

(Continued from page 46A)

Wysocki, K., Hazeltine Electronics Corp., 58–25 Little Neck Pkwy., Little Neck, L.I., N. Y.

Admission to Member

Abbott, W., Engineering Annex, Ames, Iowa
Anderson, H. J., 1640 Wisconsin Ave., Flint 6, Mich.
Borden, C. C. Sr., 33 Clinton Ave., New Providence, N. J.
Bradley, J. C., 289 Carmita Ave., Rutherford, N. J.
Britney, O. L., Rm. 244, Hunter Bldg., Ottawa, Ont., Canada
Burgess, N. M., 15 Browning Ave., Toronto, Ont., Canada
Chaney, H. L., Box 724, Red Bank, N. J.
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(Continued on page 50A)

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Filament Potential	6.0 ± 5% ac or dc volts
Filament Current	0.80 amp.
Plate Potential	450 max. dc volts
Screen Potential	250 max. dc volts
Grid Potential	–125 max. dc volts
Plate Current	75 max. dc ma.
Plate Dissipation	15 max. watts
Screen Dissipation	4 max. watts
Grid Driving Power (Class C)	0.5 watt approx.
Power Output (Class C)	20 watts

AVERAGE DIRECT INTERELECTRODE CAPACITANCES

Grid to Plate (with external shielding)	0.18 max. mmfd
Input	8.5 mmfd
Output	6.0 mmfd

MECHANICAL DATA

Maximum Overall Length	4 3/16 inches
Maximum Diameter	1 7/16 inches
Bulb	T-11
Cap	Small metal
Base	7-pin med. short shell low-loss octal

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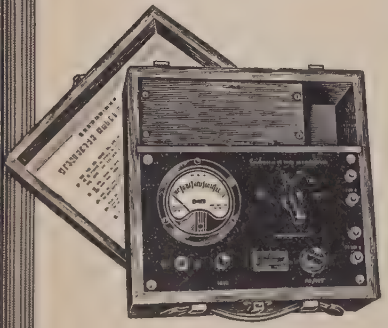

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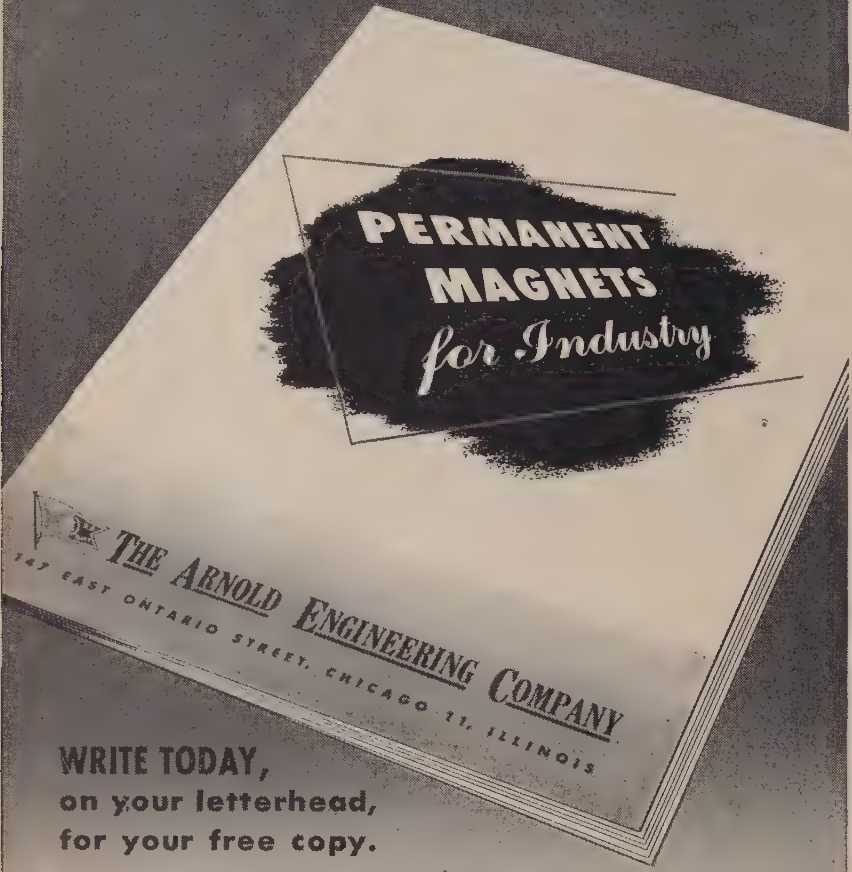
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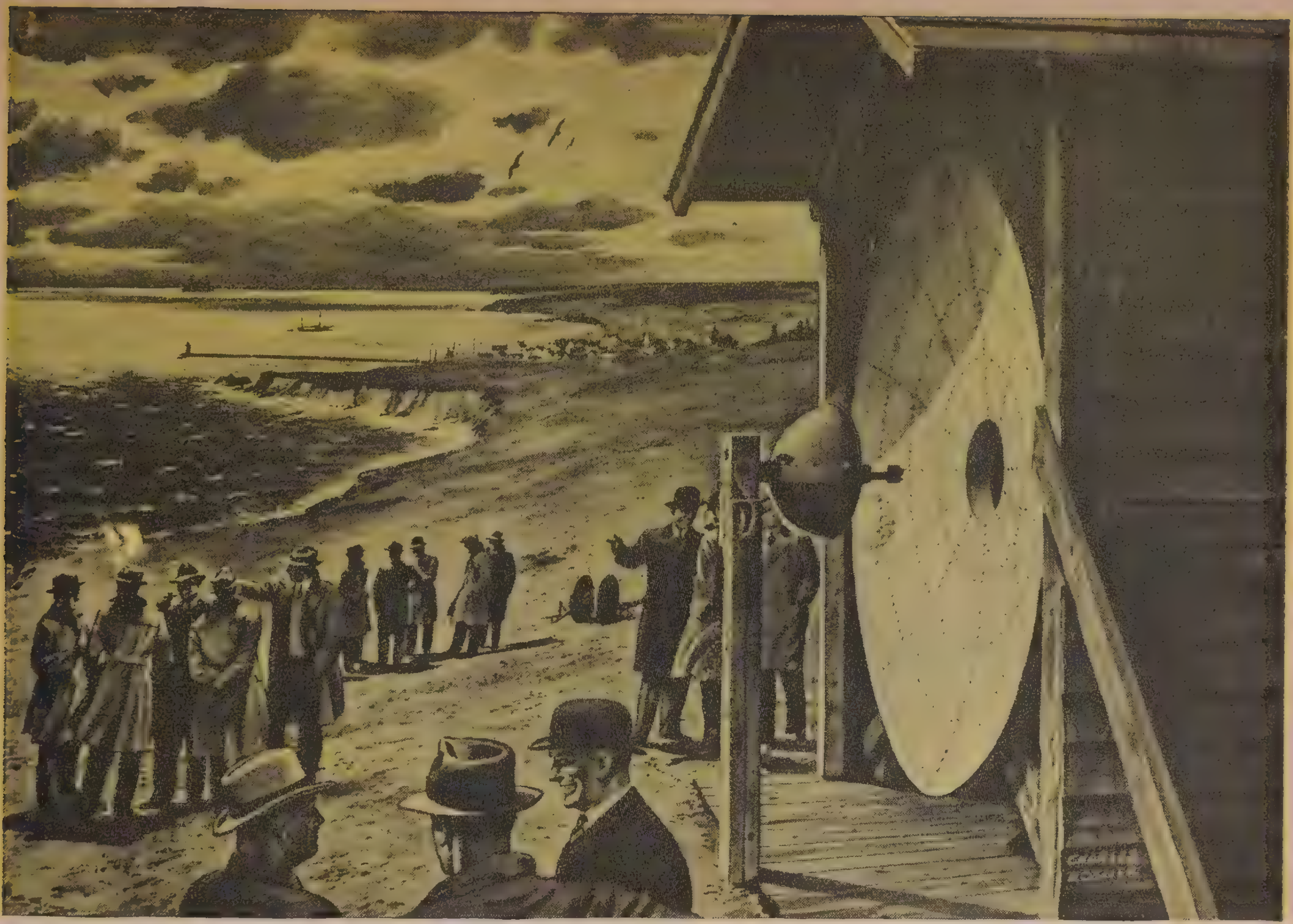

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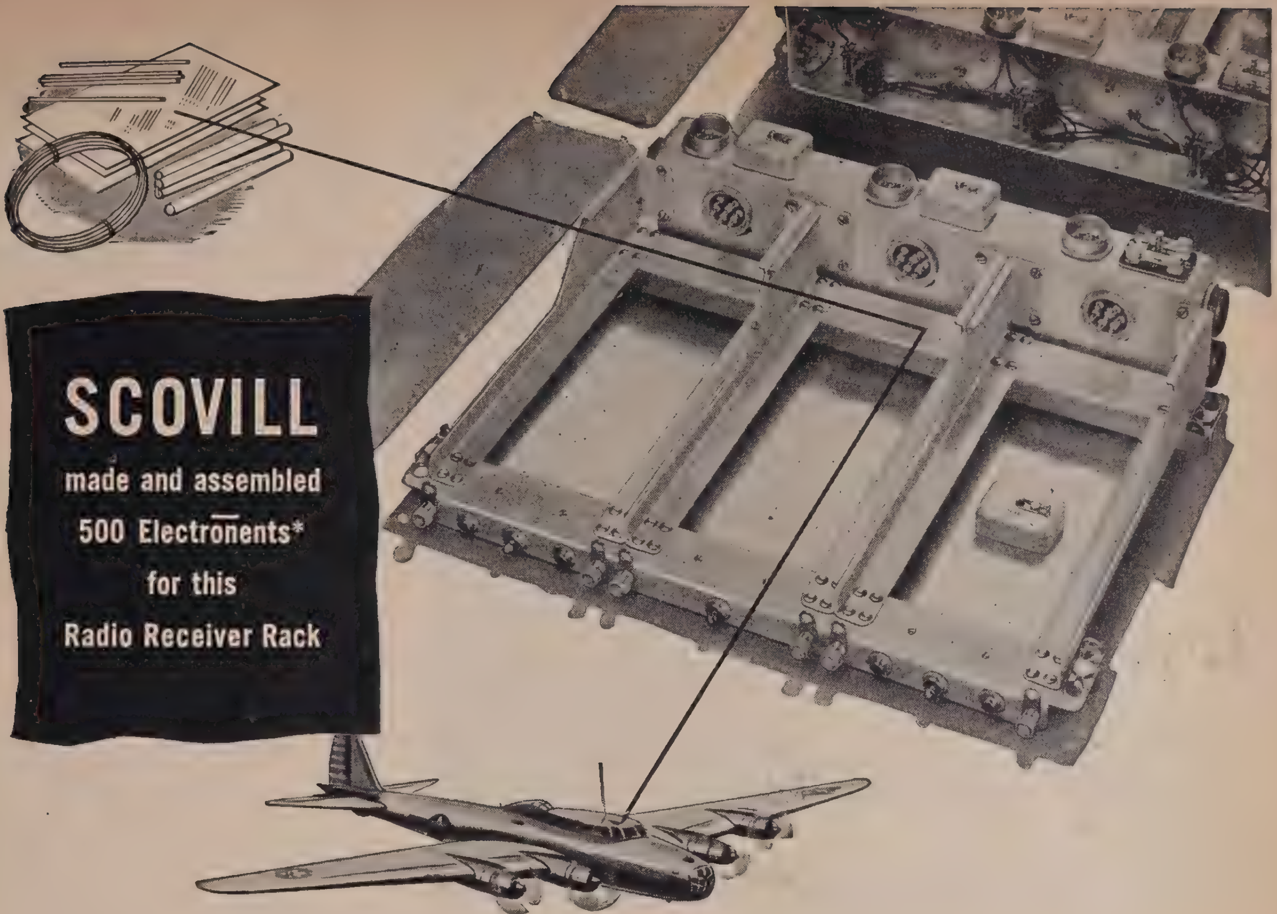

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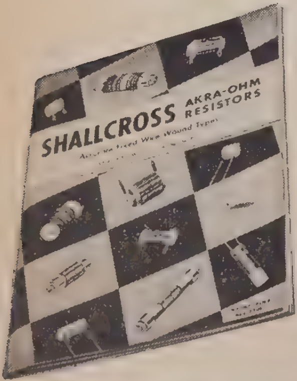

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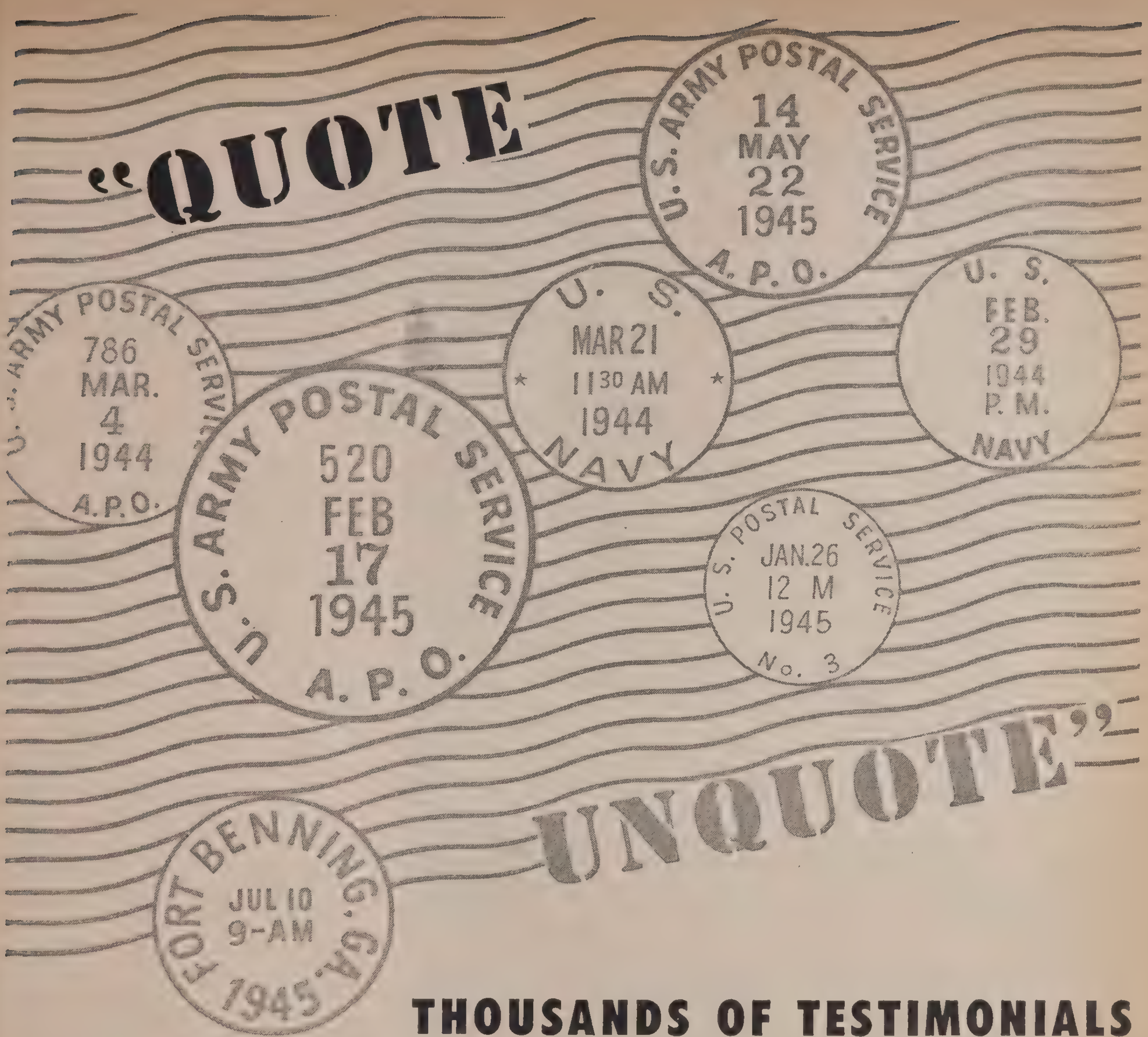
(Continued on page 42A)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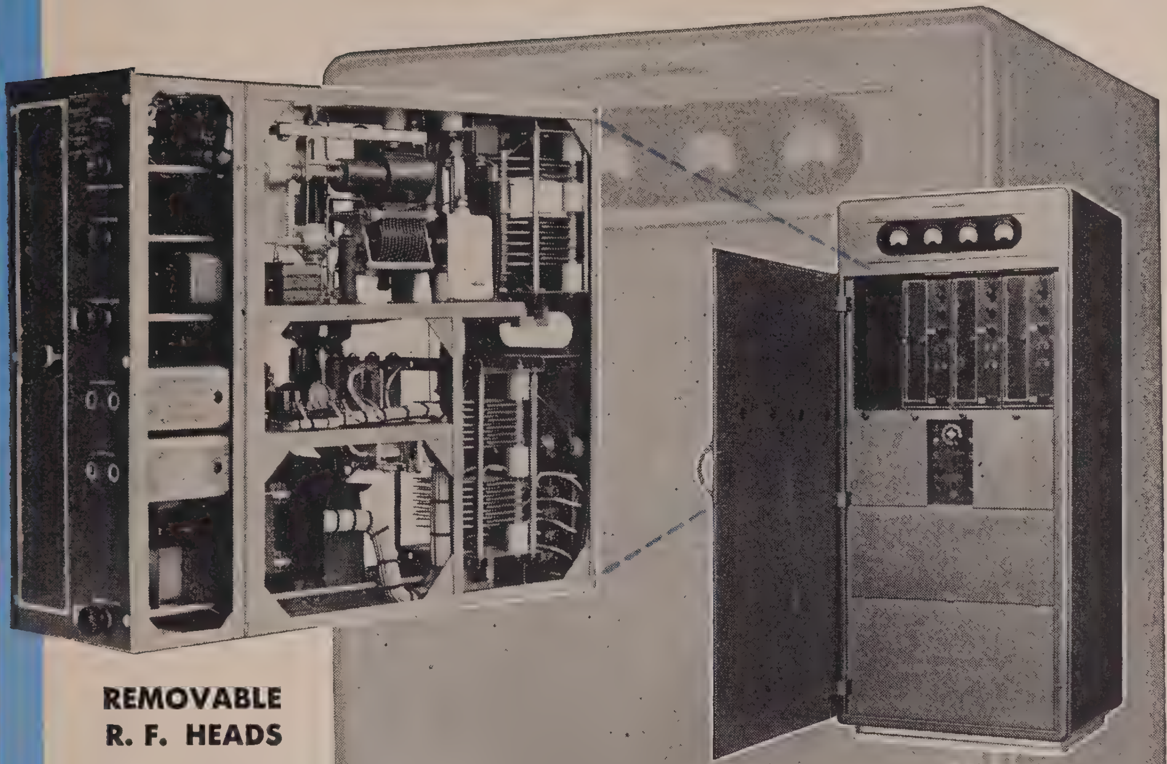
(Continued on page 46A)

Wilcox Type 99A Transmitter



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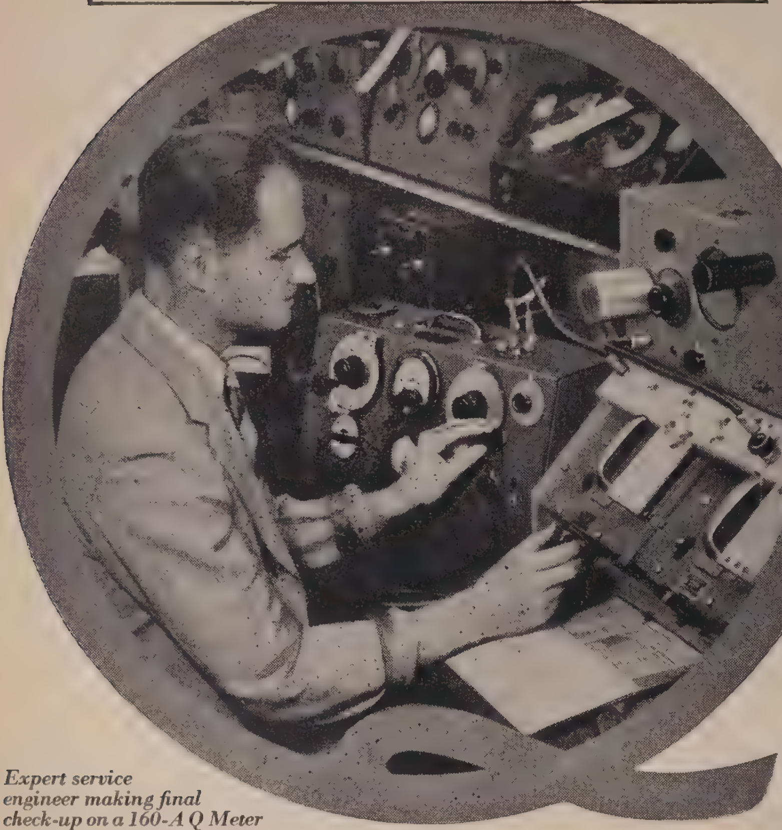

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Electronic Tubes, Rectifiers, Power Supplies, Intercommunicating Systems, Sound Systems, Photo-Cell Equipment, Batteries, Chargers, Converters, Generators, Supplies for Resistance Welders, Fuses, Test Instruments, Meters, Broadcast Station Equipment, Relays, Condensers, Capacitors, Resistors, Rheostats, Transformers, Switches, Coaxial Cable, Wire, Soldering Irons, Microphones, Speakers, Technical Books, etc.

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(Continued from page 50A)

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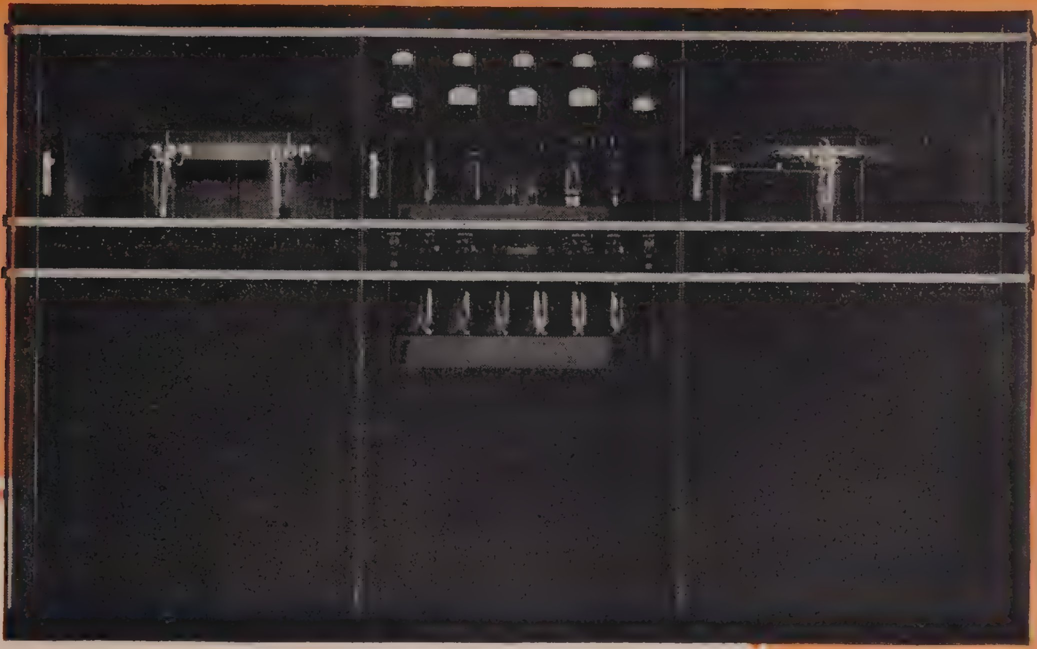
(Continued on page 60A)

The New Collins 21A, 5 kw Broadcast Transmitter



Fulfilling the Tradition of Collins Quality Leadership

The 21A is a thoroughly developed 5 kw AM broadcast transmitter, and an excellent example of characteristically superior Collins engineering and construction.

Based on sound, well-proved principles of design, the 21A has been completely modernized within recent months. New components of improved design, with longer life and higher safety factors than were previously available, assure reliable continuous operation.

The response curve is flat, within ± ½ db. from 30 to 10,000 cycles. Reduced power to 1 kw is obtained by instantaneous lowering of plate voltages, permitting uninterrupted program transmission.

We will be glad to send you detailed information regarding the 21A, other Collins transmitters, the 12Y remote amplifier, the 12Z four channel remote amplifier and Collins high quality studio equipment. Collins Radio Company, Cedar Rapids, Iowa; 11 West 42nd Street, New York 18, N. Y.

MICROPHONE
COLLINS
POWER

The Collins 12Y Remote Amplifier

A one channel remote amplifier for unattended operation from a 115 volt a.c. power source, the 12Y provides the advantages of quick set-up, small size, light weight, high fidelity, simple operation, utmost reliability and low cost. It is practically hum free due to the removal of the isolation transformer, which is in the power cable.

ARMY E NAVY

FOR BROADCAST QUALITY, IT'S . . . COLLINS

COLLINS EQUIPMENT IS SOLD IN CANADA
BY COLLINS-FISHER, LTD., MONTREAL.

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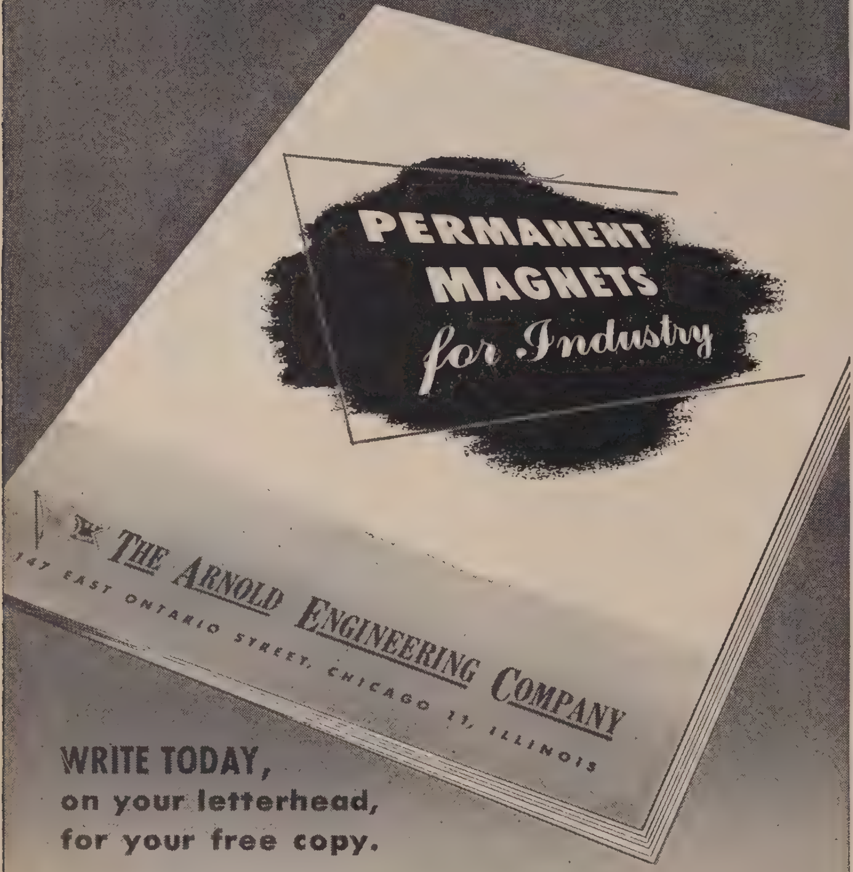
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for Industry
THE ARNOLD ENGINEERING COMPANY
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This unusual Arnold manual on permanent magnets is the product of many months of careful research and planning.

It is devoted entirely to the consideration of the factors affecting the design, fabrication and application of Alnico permanent magnets. Written entirely by Arnold engineers, its purpose is to help engineers in industry to better utilize the magnetic and physical characteristics of the Alnico alloys in arriving at efficient design.

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Specialists in the Manufacture of ALNICO PERMANENT MAGNETS

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(Continued from page 58A)

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● MODEL 200-A IMPEDANCE BRIDGE is a portable, self contained instrument of highest quality used extensively by the Army, Navy, and many manufacturers.

The range of measurement for capacitance is 1 micro microfarad to 100 microfarads; for resistance, 1 milliohm to 1 megohm; for inductance, 1 microhenry to 100 microhenrys. The accuracy on the main decade is 1% for capacitance or resistance measurements and 2% for inductance tests.

Reading obtained from 6 inch direct reading dials. All controls and connections plainly marked and conveniently located on the panel. 35-page book gives methods for many types of measurements.

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Our factory is in a position to make fast deliveries on Model 200-A and other products including precision mica condensers, binding posts, several types of AWS rheostat-potentiometers and decade and low capacity switches.

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DIAGNOSIS *by a Specialist*

THE CURE OF RADIO NOISE is a highly specialized task that involves much more than simply "hooking a condenser across the line". It requires exact knowledge of the proper size and type of capacitor to use . . . of the correct place to add it to the noise-making circuit . . . of the necessary length or positioning of connecting leads . . . and of many other seemingly trivial, but actually vital, bits of information that cannot rightfully be expected of the electrical design engineer.

This exact knowledge is available to you when you must provide radio silence for electrical apparatus. Just send us the offending equipment and we will measure its radio noise output according to standard specifications, will design the most efficient Filterette to cure the noise, will specify the proper means of installing it, and, upon your adoption of our recommendations, will authorize your use of the FILTERIZED label that tells buyers your apparatus will not interfere with radio reception. This service is free to users of Tobe Filterettes . . . write for details.

SUPPRESS MAN-MADE RADIO STATIC
TOBE
FILTERIZED

TOBE DEUTSCHMANN CORPORATION-CANTON, MASSACHUSETTS

ORIGINATORS OF FILTERETTES . . . THE ACCEPTED CURE FOR RADIO NOISE

Tests Prove Eimac Vacuum Condensers Far Superior in Operating Efficiency

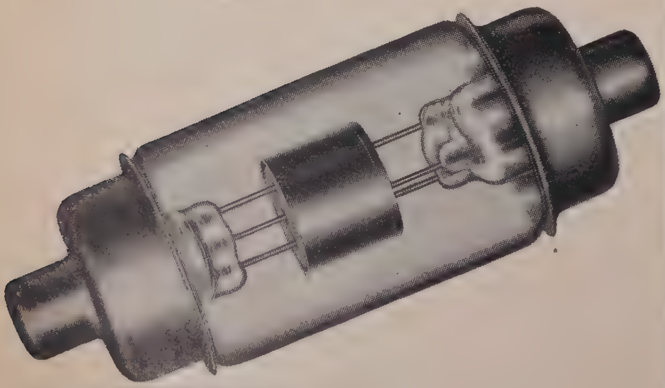
Ability to handle high current at high frequencies is the true measure of the performance of a capacitor. A high peak voltage rating based on low frequency measurements does not tell the whole story.

The chart on this page shows the results of tests at 50 Mc. conducted on a standard Eimac VC50-32 Vacuum Capacitor and three other 50 mmfd. vacuum capacitors, designated on the chart by "A," "B" and "C." At just over 17 amps. (approximately 1525 peak volts across the capacitor) Unit "A" (rated at many times the applied voltage) became sufficiently heated to melt the solder on the end caps. Under this same test, the Eimac VC50-32 operates at less than 70°.

Eimac introduced the vacuum capacitor in 1938. It is interesting to note that the original Eimac capacitor design is still outperforming all comers. Such outstanding performance is typical of all Eimac products, which is one of the reasons why they are first choice of leading electronic engineers throughout the world.

1113

Follow the leaders to **Eimac** REG. U.S. PAT. OFF. **TUBES**



EIMAC VACUUM CAPACITOR TYPE VC50-32
General Characteristics

Mechanical:

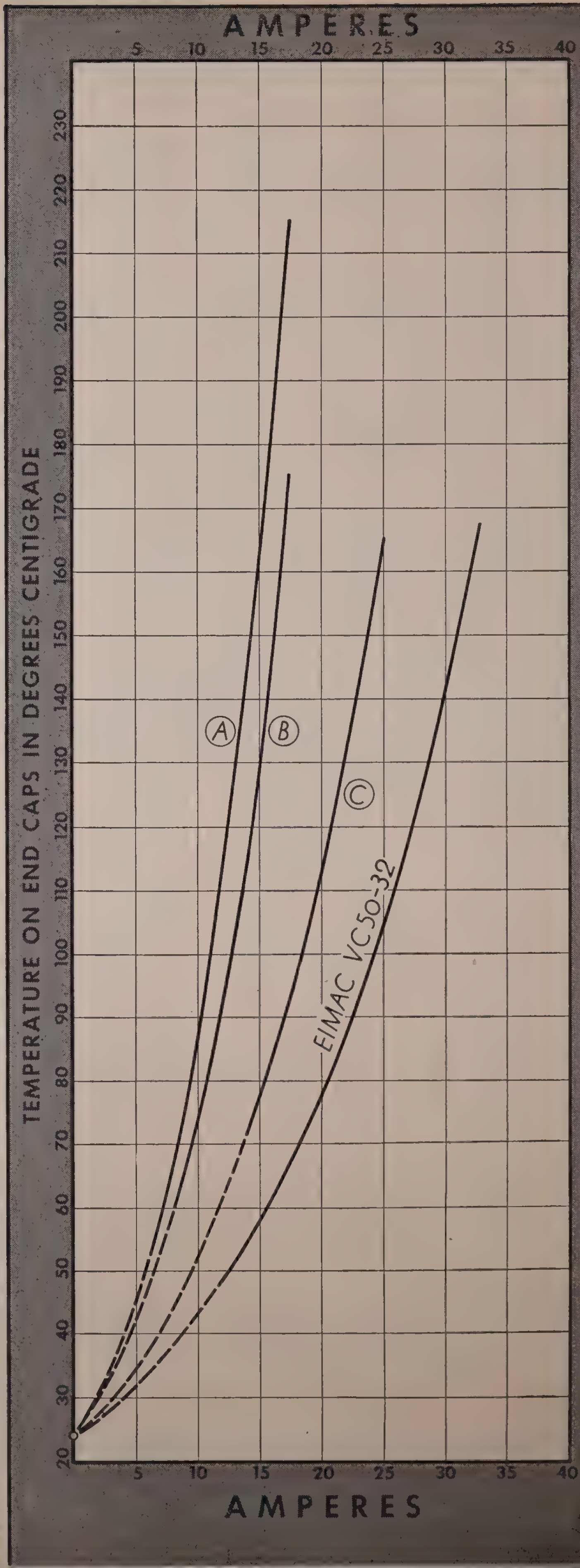
Maximum Overall Dimensions

Length	6.531 inches
Diameter	2.281 inches

Electrical:

Maximum Peak Voltage	32,000 volts
Maximum RMS Current	28 amps.

EITEL-McCULLOUGH, INC., 1113 San Mateo Avenue, San Bruno, Calif.
Plants located at: San Bruno, California and Salt Lake City, Utah
Export Agents: Frazar & Hansen, 301 Clay St., San Francisco 11, Calif., U. S. A.



We grew up with electronics

Our engineers and executives grew up with Electronics. Before the war we manufactured commercial radio equipment. During the war we greatly expanded our engineering and research staff and did extensive work in advanced electronics for the Army and Navy. Our present engineering and research facilities occupy more than 30,000 square feet of space. Our current production program is centered on communications equipment for rail, air, highway, marine and commercial use. Other products, notably in the field of industrial electronics, are under development. Aireon's engineering and research staff will be glad to consult with you on your electronic problems. Your inquiry will have prompt attention.

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Radio and Electronics • Engineered Power Controls

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PROCEEDINGS OF THE I.R.E., December, 1945, vol. 33, no. 12. Published monthly by The Institute of Radio Engineers, Inc., 330 West 42nd St., New York 18, N.Y. Price, $1.00 per copy. Subscriptions: United States and Canada, $10.00 a year; foreign countries, $11.00 a year. Entered as second-class matter October 26, 1927 at the post office at Menasha, Wisconsin, under the act of February 28, 1925. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Measuring SOUND

In flight tests at Wright Aeronautical, a Western Electric sound analyzer is used to measure sound characteristics of the plane and locate major sound disturbances.

Why Western Electric equipment leads the way!

1. **Western Electric products are designed by Bell Telephone Laboratories—world's largest organization devoted exclusively to research and development in all phases of electrical communication.**

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3. **The outstanding quality of Western Electric equipment has been proved daily on land, at sea, in the air, under every extreme of climate. No other company supplied so much equipment of so many different kinds for military communications.**

Western Electric

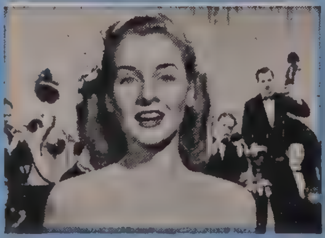
Today's world is a world of sound. How different it would be without the telephone, radio, public address systems, aids for the hard of hearing, talking pictures!

For many years, Bell Telephone Laboratories and Western Electric—working closely as research and manufacturing teammates—have led the way in building this world of sound.

In the course of their sound-transmission work, these teammates

Western Electric has specialized

AM • BROADCASTING • FM



TELEVISION

AVIATION RADIO

MARINE RADIO

or spreading it around

A powerful Western Electric public address system spreads sound evenly throughout New York's huge Madison Square Garden which seats more than 18,000 people.

equipment leads the way!

have also developed scientifically accurate instruments for *measuring* and *analyzing* sound and vibration. These instruments have many important uses today—will have still more tomorrow.

Through their lifetime of pioneering in this field, Bell Labs and Western Electric have gained a unique knowledge of sound and how to handle it. Count on them for the finest equipment for measuring sound or spreading it around!

Western Electric

Distributed by Graybar Electric Company

Buy all the Victory Bonds you can
. . . and keep all you buy!

Knowledge in all of these fields

MOBILE RADIO

HEARING AIDS

SOUND MOTION PICTURES

VACUUM TUBES

I. R. E. Winter Technical Meeting and Radio Engineering Show

January 23 to 26, 1946

Hotel Astor, New York City

I.R.E. MEETING
JAN. 23 to 26, 1946
HOTEL ASTOR

More than 3,000 members attended the last I. R. E. winter meeting—and with travel restrictions lifted, an even greater attendance is expected in 1946. Don't fail to make your reservations now for three days devoted to interesting and instructive technical papers—plus unusual features and entertainment.

TECHNICAL PAPERS

The lid has been lifted! Vital papers held back for security reasons will be among those presented in two and one-half days of technical sessions.

EXHIBITS

Two floors of the Hotel Astor will be given over to 150 exhibits revealing wartime advances and post-war equipment. War developments applicable to civilian equipment will be feature attractions.

BANQUET

The annual banquet, on January 24, is the social highlight of the I.R.E. year. 2,500 feasters will hear a nationally prominent speaker, see two major awards made, be entertained royally.

WOMEN'S PROGRAM

There'll be plenty to keep the better half busy—trips to points of interest, entertainment—and no men.

PRESIDENT'S LUNCHEON

The incoming president will be honored on January 25, at a get-together which has come to be a feature of these annual meetings.

DON'T DO THIS!

USE THIS →

The hotel situation in New York is still tight — and the parks are cold in January! Fill in the coupon to the right and mail it today.

The Institute of Radio Engineers
Advertising Department
303 West 42nd Street, Room 707
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Gentlemen:

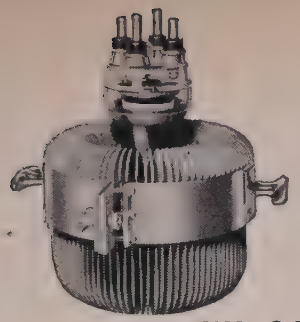
Please send me an application form for reservations and a list of cooperating hotels.

Name ____________________

Address ____________________

Organization ____________________

The demand for hotel accommodations is so critical that reservations for the Winter Technical Meeting are being handled by the New York Convention and Visitors Bureau, a cooperative civic organization. Arrangements are being made to accommodate the membership in several New York hotels. Mail this coupon *immediately*.


WL-889R

WL-889

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OSCILLATING

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WESTINGHOUSE

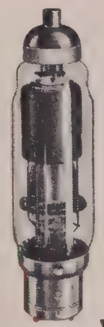
ELECTRONIC TUBES FOR ELECTRONIC HEATING

Westinghouse manufactures a complete line of electronic tubes that will meet your RF heating requirements. For descriptive data on any of the types shown, call your local Westinghouse district office or write:

Electronic Tube Sales Department,
Westinghouse Electric Corporation,
Bloomfield, New Jersey

WL-892R

WL-677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CONTROLLING

WL-857B

WL-575A

WL-872A/872

RECTIFYING

TUNE IN: John Charles Thomas
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Westinghouse

Electronic Tubes at Work

Here's the easiest way to get 10 WATTS OUTPUT at 156 Mc!

Build your Xmtr around HK-24G *Gammatrons*

JAN 3C24

Designing transmitters for the 152 to 162 megacycle region can be greatly simplified through the use of HK-24Gs. In the circuit below one of these tubes produces 10 watts at 156 Mc; more than ample to excite a pair of HK-257Bs in a 400 watt final.

These efficient Gammatrons, with low inter-electrode capacities and short plate and grid leads, are the answer for those who are engineering 1.9-meter equipment for railroads, police point-to-point, fire, press or broadcast relays, maritime, geophysical, urban telephone, and experimental use. They are also ideally suited for amateur transmitters on the 144-148 megacycle band, authorized as of November 15.

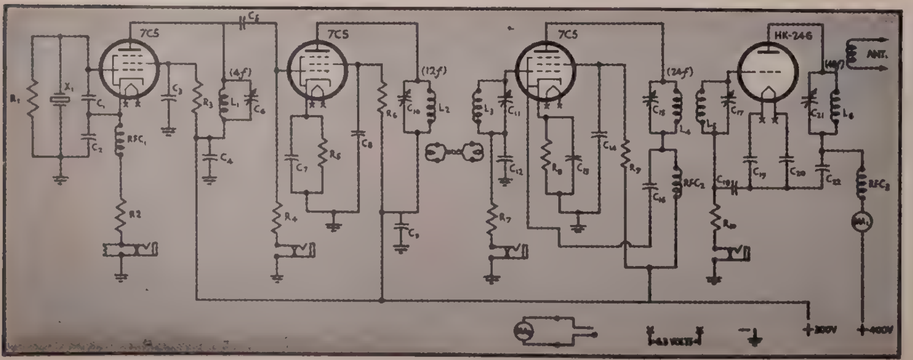
Not only will HK-24Gs do the job, but they will do it at low cost. Made with tantalum elements, without internal insulators, they can withstand heavy overloads and have a long operating life. And they are offered at a new low price made possible by improved production methods.

HK-24Gs are available now in quantity . . . and are one of Heintz and Kaufman's *standardized* types. Maximum plate dissipation 25 watts. Write today for data.

Now

ONLY $6.00

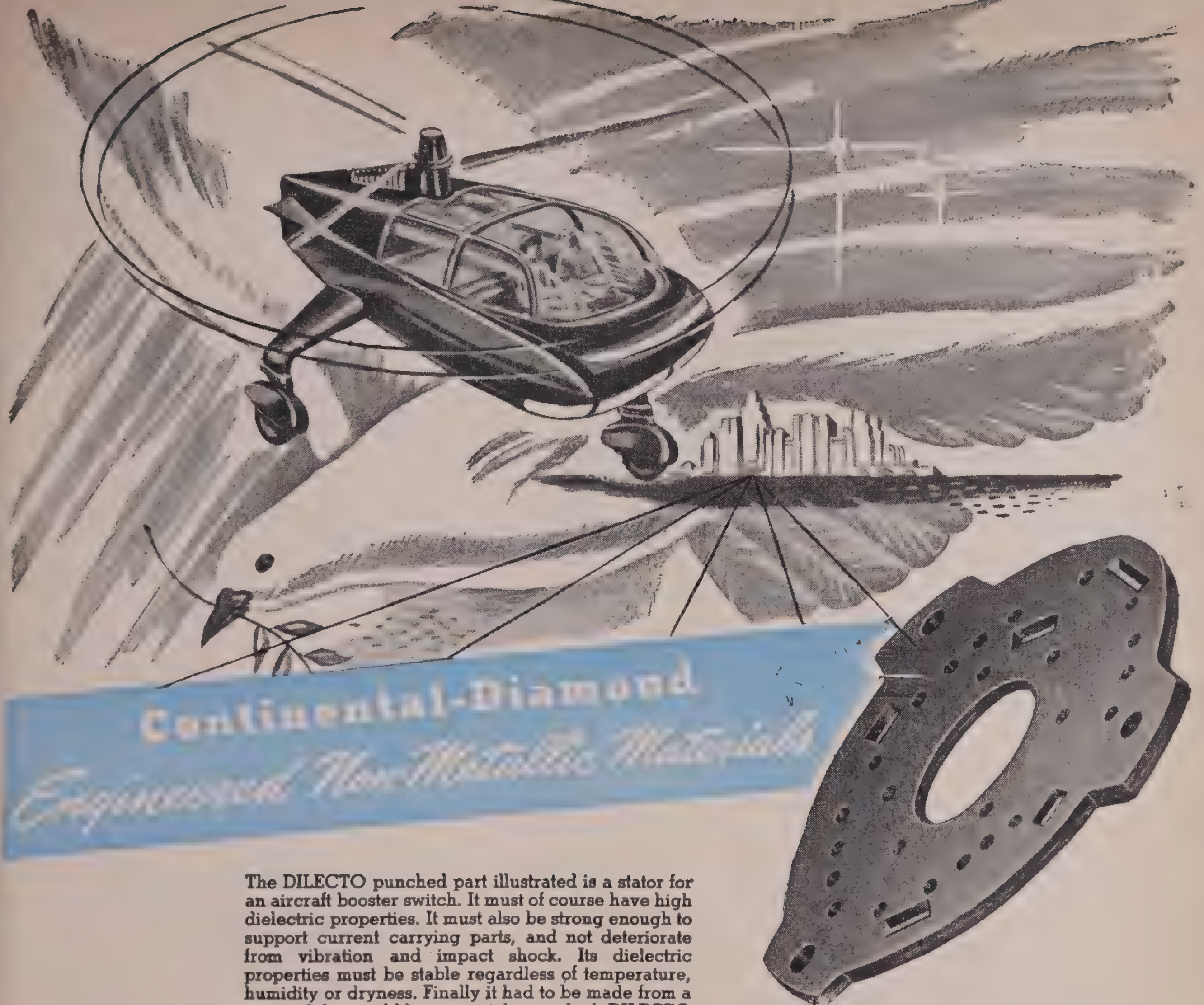
HEINTZ AND KAUFMAN LTD.
SOUTH SAN FRANCISCO · CALIFORNIA
EXPORT AGENTS: M. SIMON & SON CO., INC., 25 WARREN STREET, NEW YORK CITY, N. Y.



COMPONENT PARTS FOR WIRING DIAGRAM

C_1—35 mmfd.
C_2—250 mmfd.
C_3, C_4, C_7, C_8, C_9, C_{12}, C_{13}, C_{14}—.002 mfd.
C_5—50 mmfd.
C_6—APC 100
C_{10}, C_{11}, C_{15}, C_{17}—APC 25 mmfd.
C_{16}, C_{18}, C_{22}—100 mmfd.
C_{19}, C_{20}—250 mmfd.
C_{21}—15 mmfd.
R_1—100,000 ohms, ½ watt
R_2—200 ohms, 1 watt
R_3—10,000 ohms, 1 watt
R_4—250,000 ohms, ½ watt
R_5, R_8—500 ohms, 1 watt
R_6, R_9—20,000 ohms, 1 watt
R_7—100,000 ohms, ½ watt
R_{10}—15,000 ohms, 2 watts
RFC_1—2.5 mh.
RFC_2,
RFC_3—Hf choke, approx. 50 turns No. 32 on ¼" form.
MA_1—0-100 ma.
MA_2—0-25 ma.
X_1—3.5 Mc. crystal
L_1—19 turns No. 22 DCC ⅜" form
L_2—7 turns No. 14 enamel, 3/16" dia. Length ¼" air, link
L_3 coupled to L_2 with one turn.
L_4—6 turns No. 14 enamel, dia. ⅜", length 7/16" air, induc-
L_5 tively coupled on same axis.
L_6—4 turns No. 12 copper, dia. ⅜", length ⅜"

Note: Grid and plate by-passes on driver and doubler final should be returned directly to tube cathode. All jacks are closed circuit jacks.



The DILECTO punched part illustrated is a stator for an aircraft booster switch. It must of course have high dielectric properties. It must also be strong enough to support current carrying parts, and not deteriorate from vibration and impact shock. Its dielectric properties must be stable regardless of temperature, humidity or dryness. Finally it had to be made from a material that could be accurately punched. DILECTO met all these requirements with a wide margin of safety.

There are many grades of DILECTO. Each developed to meet specific electrical, mechanical, chemical or thermal problems. Special grades can be developed to meet unusual problems. DILECTO is also available in combination with Diamond Fibre to still further enlarge its sphere of usefulness. This C-D NON-metallic may be the answer to your "What Material?" problem, in your present and future products, whether used in the air, on land or sea.

KE-45

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The Plastics

DILECTO—A Laminated Phenolic.
CELORON—A Molded Phenolic.
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Available in Standard Sheets, Rods and Tubes; and Parts Fabricated, Formed or Molded to Specifications.

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Bulletin GF gives Comprehensive Data on all C-D Products. Individual Catalogs are also Available.

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Established 1895.. Manufacturers of Laminated Plastics since 1911—NEWARK 48 • DELAWARE

High power output, long life, feature these transmitting tube stalwarts!

- **TYPE GL-892**
  Water-cooled . . . $170
- **TYPE GL-892-R**
  Forced-air-cooled . $345

HERE is *proved* power, dependability, and long service life for the large AM transmitter owner or the manufacturer using electronic heating. General Electric Types GL-892 and GL-892-R have demonstrated their reliability in broadcasting and industrial sockets operating 24 hours a day, 7 days a week. With broad applications as high-power amplifiers, modulators, and oscillators, Types GL-892 and GL-892-R also are adaptable as to filament supply, their 2-unit filament permitting operation from 2-phase or single-phase a-c, as well as from d-c. For complete data to supplement the basic ratings at the right, see your nearest G-E Office or distributor, or write *Electronics Department, General Electric Company, Schenectady, N. Y.*

Come to GENERAL ELECTRIC with any special tube problems. A staff of experienced G-E tube engineers will work with you closely to meet your application or replacement needs.

CHARACTERISTICS

Three-electrode high-vacuum power tubes for use as amplifiers and modulators in broadcasting and communications equipment—also oscillators in industrial electronic heating. Besides Types GL-892 and GL-892-R shown above, Types GL-891 and GL-891-R also are available at the same prices, and are similar in design characteristics except for the amplification factor, as given below.

Rating	GL-892	GL-892-R	GL-891	GL-891-R
Filament voltage	11 v	11v	11v	11 v
Filament current	60 amp	60 amp	60 amp	60 amp
Max plate voltage	15,000 v	12,500 v	12,000 v	10,000 v
Max plate current	2 amp	2 amp	2 amp	2 amp
Max plate input	30 kw	18 kw	18 kw	15 kw
Max plate dissipation	10 kw	4 kw	6 kw	4 kw
Amplification factor	50	50	8	8

Notes: (1) Filament voltage and current given above, are per unit of 2-unit filament. (2) Maximum frequency for all four tube types is 1.6 megacycles at max plate input; up to 20 megacycles at reduced ratings.

TRANSMITTING, RECEIVING, INDUSTRIAL, SPECIAL PURPOSE TUBES · VACUUM SWITCHES AND CAPACITORS

GENERAL ELECTRIC

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ALSIMAG
TRADE MARK REGISTERED U. S. PATENT OFFICE

CERAMICS

SO RUGGED THEY COULD WITHSTAND THE SHOCK OF BEING FIRED FROM A GUN WITH A FORCE OF

20,000g

in the

'RADIO PROXIMITY FUSE'

War's Number 2 Scientific Development

ALSIMAG Ceramic Insulators were extensively used in condensers for the 'Radio Proximity Fuse' described by high Navy officials as second only to the atomic bomb among the greatest scientific developments of the war.

Development of the fuse required production of electronic parts so rugged they could withstand the shock of being fired from a gun with a force 20,000 times that of gravity. The components had to be so small that a complete unit could be installed in the nose of a projectile.

The fuse, developed at a cost of $800,000,000 is an extremely rugged, five tube radio sending and receiving station which fits into the nose of a projectile. Reflected impulses explode the projectile when it passes within 70 feet of enemy planes.

The 'Radio Proximity Fuse' was the effective answer to Japanese suicide plane attacks, as well as buzz bomb attacks on London.

American Lava Corporation is justly proud of the fact that it was able to provide the Ceramic Insulators capable of withstanding the tremendous shock of being fired from a gun in the 'Radio Proximity Fuse.'

Whatever you are planning in the field of electronics, we believe our specialized knowledge, research and production facilities will prove helpful. Let's work together.

ALCO has been awarded for the fifth time the Army-Navy "E" Award for continued excellence in quantity and

AMERICAN LAVA CORPORATION
CHATTANOOGA 5, TENNESSEE

43RD YEAR OF CERAMIC LEADERSHIP

more efficient
...in miniature
ACTUAL SIZE

The cast iron pump was modern two or three generations ago. It was a big improvement over the old oaken bucket. But today we use a comparatively small faucet that supplies water at a twist of the wrist. It is another milestone on the road to greater efficiency in miniature.

This same tendency is evident in the development of the Electronic Tube. The Tung-Sol Miniature is the result of the trend to smaller component parts. It is used to great advantage in reducing the over-all size of equipment. But more important, Tung-Sol Miniatures do a more efficient job than the old style tube; especially in high frequency circuits. They have a low capacity and high mutual conductance. Shorter leads give them low inductance. Smaller elements weigh less, making Miniatures more rigid. This helps to eliminate distortion from vibration.

When planning new electronic devices or when improving old ones, discuss circuits and tube selection with Tung-Sol engineers. Their services are at your disposal. Such conferences are held in strictest confidence.

TUNG-SOL

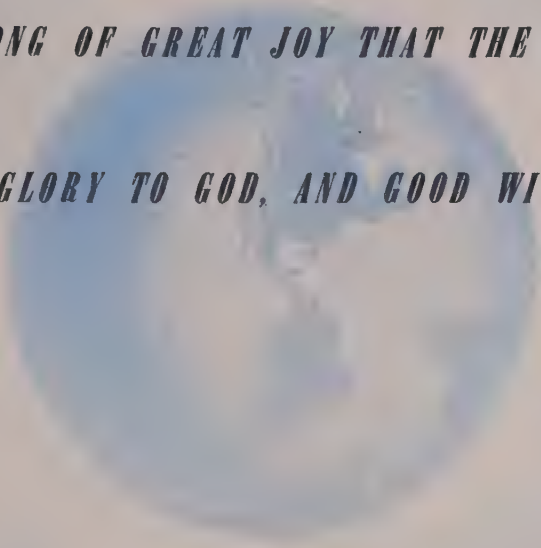
vibration-tested

ELECTRONIC TUBES

TUNG-SOL LAMP WORKS INC., NEWARK 4, NEW JERSEY
Also Manufacturers of Miniature Incandescent Lamps, All-Glass Sealed Beam Headlight Lamps and Current Intermittors

BLOW, BUGLES OF BATTLE, YOUR MARCHES OF PEACE;
EAST, WEST, NORTH AND SOUTH, LET THE LONG QUARREL CEASE.
SING THE SONG OF GREAT JOY THAT THE ANGELS BEGAN,
SONG OF GLORY TO GOD, AND GOOD WILL TO MAN..

WHITTIER



As the tumult and the shouting die . . . and the Yuletide bells ring out their old, old story of peace on earth . . . it is good to make merry and wish our fellowmen all over the world . . . Merry Christmas and a Grand New Year . .

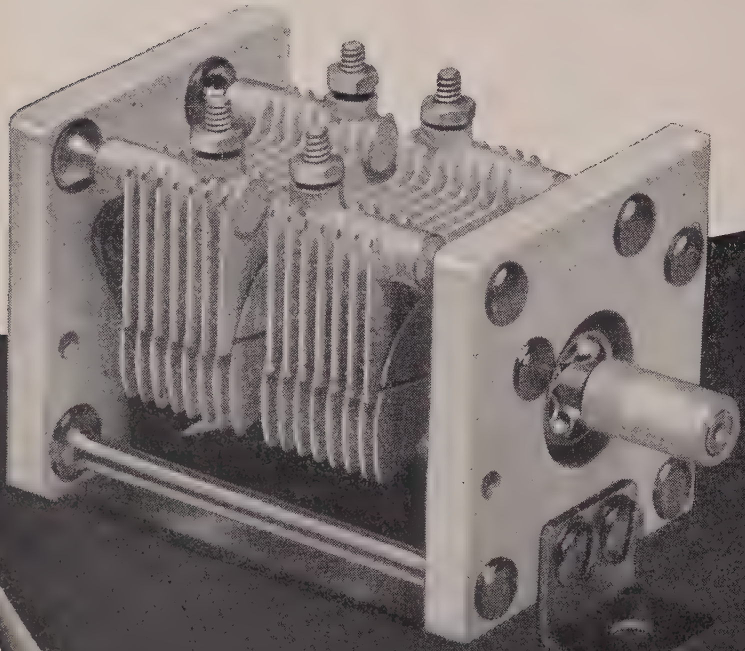
GENERAL INSTRUMENT CORPORATION
829 NEWARK AVENUE • ELIZABETH 3, N. J.

VARIABLE CONDENSERS • TUNING MECHANISMS • RECORD CHANGERS • SPEAKERS • OTHER RADIO COMPONENTS

Copyright 1945, G. I. Corp. [illegible]

new VHF-UHF

CAPACITOR TECHNIQUE



Send card for mechanical and electrical specifications.

Here's the answer to your UHF design problems. Noiseless operation — no rotor contacts — symmetrical layout. The new "VU" type variable capacitors can be used in conventional tuned circuits at frequencies as high as 500 megacycles. Write for folder with full technical data.

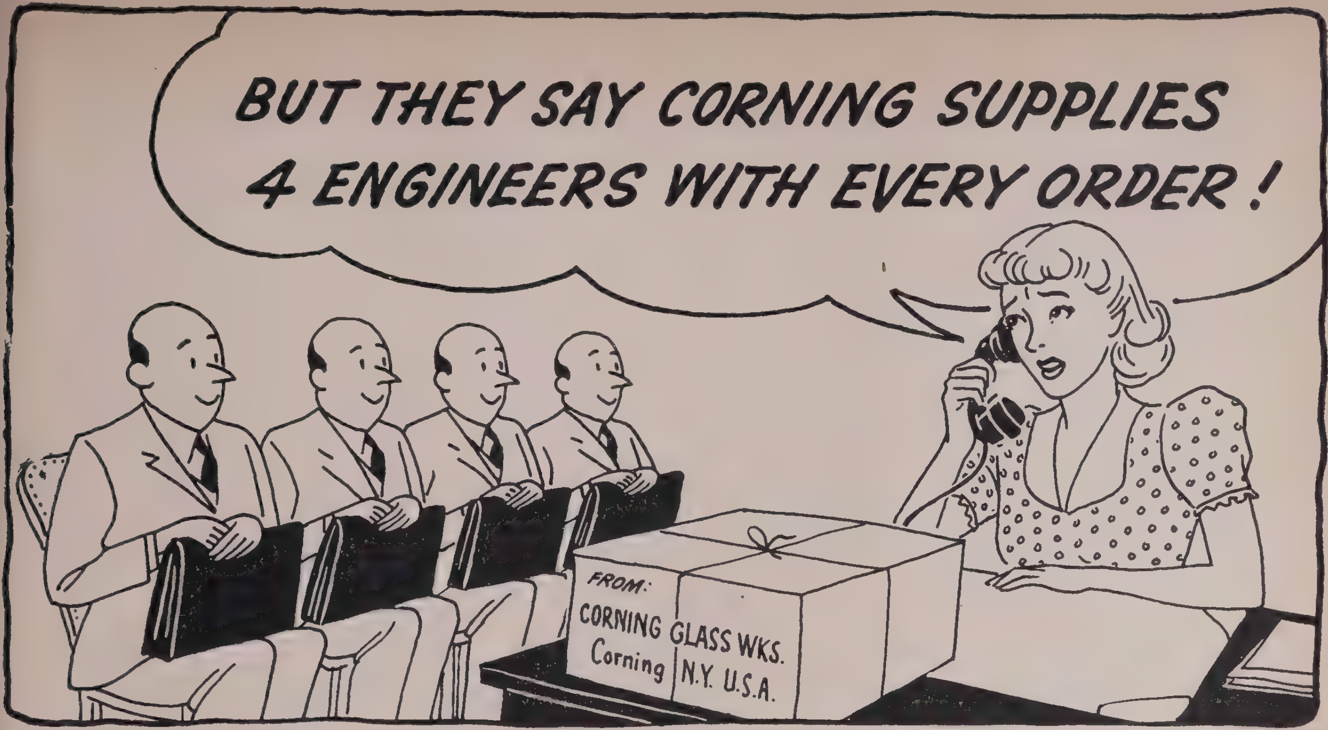
THE SIGN OF QUALITY

ESTABLISHED 1910

HAMMARLUND

THE HAMMARLUND MFG. CO., INC., 460 W. 34TH ST., NEW YORK 1, N.Y.

MANUFACTURERS OF PRECISION COMMUNICATIONS EQUIPMENT


BUT THEY SAY CORNING SUPPLIES 4 ENGINEERS WITH EVERY ORDER !
FROM:
CORNING GLASS WKS.
Corning N.Y. U.S.A.

HOW 4 SPECIAL CORNING SERVICES CAN SAVE YOU LOTS OF GRIEF!

IT'S quite a job getting a new electronic product into production. Materials, methods and prices buzz around your head like a bunch of bees. But you don't have to solve your problems all alone. For Corning has four special engineering services to help you:

1. Sales Engineers—To keep you in touch with latest developments and explain your problems to Corning's technical experts for prompt solution.

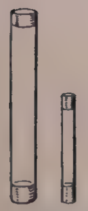
2. Product Engineers—Technical men who translate Corning Research in Glass into practical applications which may solve your particular headaches.

3. Plant Engineers—These men are anxious to see you get the best possible price on your order. They often point out changes in design which reduce costs.

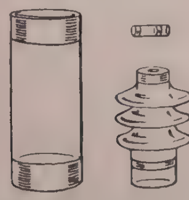
4. Technical Service Engineers — These men get you started right. They help your people lick the production bugs.

Of course, Corning Electronic Glassware also means thousands of glass formulae so you can get the right one for your job. It means Corning's unique metallizing process forming a permanent bond between glass and metal. Tubes, bushings, headers, etc., can be soldered in place to form permanent hermetic seals. It means an entire plant at Bradford, Pa., devoted exclusively to the manufacture of electronic specialties quickly, in large quantities. To get the fastest service in solving your pet problem, write, wire or phone Electronic Sales Department, P-12, Technical Products Division, Corning Glass Works, Corning, New York.

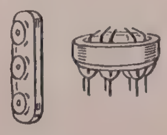
Note *— The metallized Tubes and Bushings, Headers and Coil Forms below are all made by the famous Corning Metallizing Process. Can be soldered into place to form true and permanent hermetic seals. Impervious to dust, moisture and corrosion.*



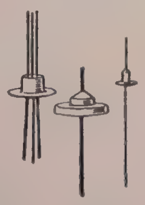
Metallized Tubes for resistors, capacitors, etc. 20 standard sizes ½" x 2" to 1¼" x 10". Mass-produced for immediate shipment.



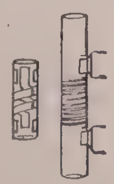
Metallized Bushings. Tubes in 10 standard sizes, $\frac{5}{16}$" x $\frac{35}{32}$" to 1" x 4$\frac{7}{32}$" in mass production for immediate shipment.



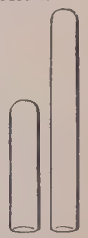
Headers—The best way to get a large number of leads in a small space for assembly in one operation.



Eyelet Terminals — Single or multiple eyelets permit design flexibility. Standard items readily available in quantity.



Coil Forms—Grooved for ordinary frequencies—metallized for high frequencies. In various designs and mountings.



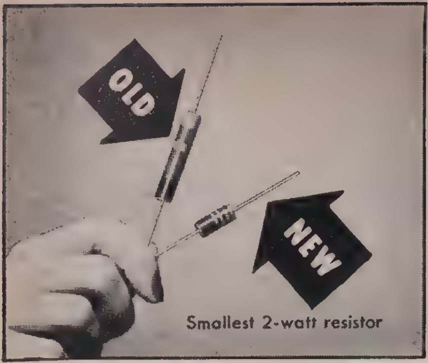
VYCOR Brand cylinders—very low loss characteristics. Stands thermal shock up to 900°C. Can be metallized.

CORNING
means
Research in Glass

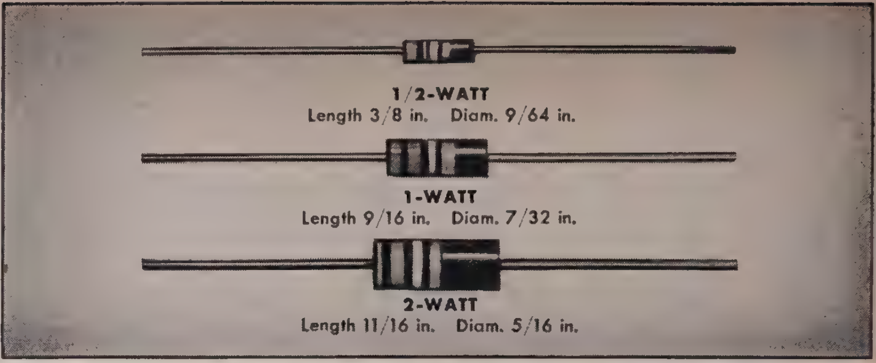
Electronic Glassware

VYCOR CORNING PYREX

"PYREX", "VYCOR" and "CORNING" are registered trade-marks and indicate manufacture by Corning Glass Works, Corning, N. Y.



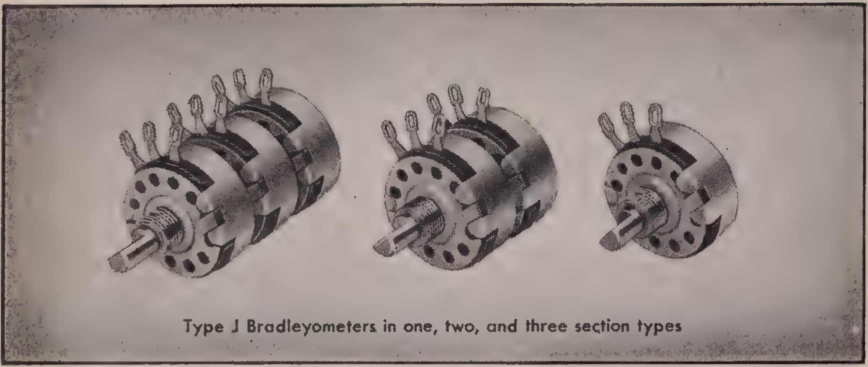
Smallest 2-watt resistor



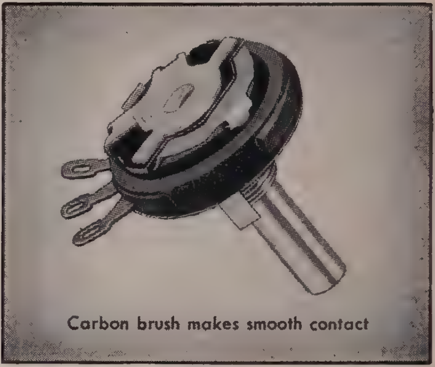
1/2-WATT
Length 3/8 in. Diam. 9/64 in.

1-WATT
Length 9/16 in. Diam. 7/32 in.

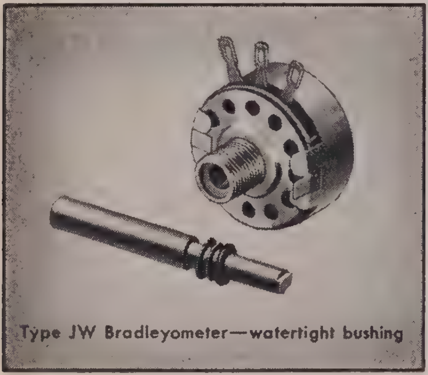
2-WATT
Length 11/16 in. Diam. 5/16 in.



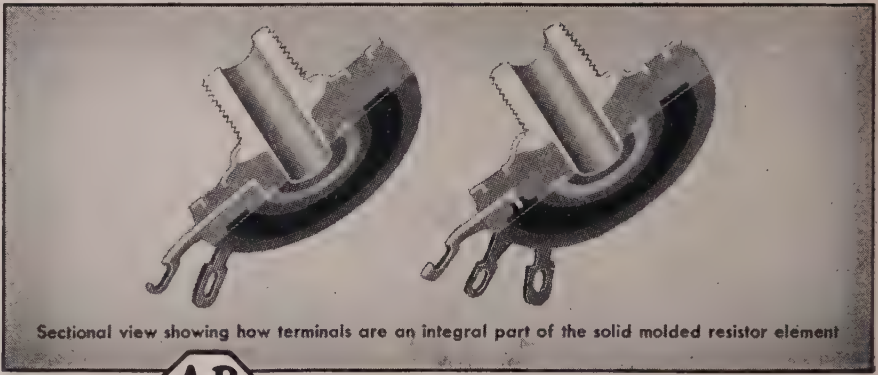
Type J Bradleyometers in one, two, and three section types



Carbon brush makes smooth contact



Type JW Bradleyometer—watertight bushing



Sectional view showing how terminals are an integral part of the solid molded resistor element

A·B
Quality

ALLEN-BRADLEY

FIXED & ADJUSTABLE RADIO RESISTORS

QUALITY

FIXED INSULATED RESISTORS—Bradleyunits are available in 1/2-watt, 1-watt, and 2-watt ratings. They will sustain an overload of ten times rating for several minutes without failing. Wax impregnation is not necessary to pass salt water immersion test. The 1/2-watt and 1-watt units are available in all RMA standard values from 10 ohms to 10 megohms. Two-watt units available from 10 ohms to 1 megohm.

ADJUSTABLE RESISTORS—Type J Bradleyometers are the only continuously adjustable composition resistors having a 2-watt rating with good safety factor. Resistor element is solid molded and has substantial thickness. Not a film, paint, or spray type. Molded as single unit complete with insulation, terminals, face plate, and bushing. No rivets or soldered connections. Any resistance-rotation curve can be provided.

Allen-Bradley Company, 114 W. Greenfield Ave., Milwaukee 4, Wis.

WHEN DEPENDABILITY AND PERFORMANCE ARE "MUSTS"... THE EXPERTS SPECIFY ALLEN-BRADLEY

IN TUBE MANUFACTURE ALL SMALL DETAILS ARE LARGE

TO *Federal*

From slender filament to anode block . . . *all* tube construction details, however small, are important to Federal. That is why this experienced and longtime manufacturer uses the illustrated high-magnification metallograph as part of its test equipment for checking raw material quality.

An example is the micro-photo inset. Here is shown oxide-free, high conductivity copper used for copper-to-glass seals . . . after the material has been reduced to a fine grain, non-porous structure through Federal's special metal-processing methods.

But whether copper, molybdenum or tungsten . . . they all are subjected to the same exclusive treatment and put through the same searching scrutiny . . . assurance that only the finest materials go to make up Federal tubes.

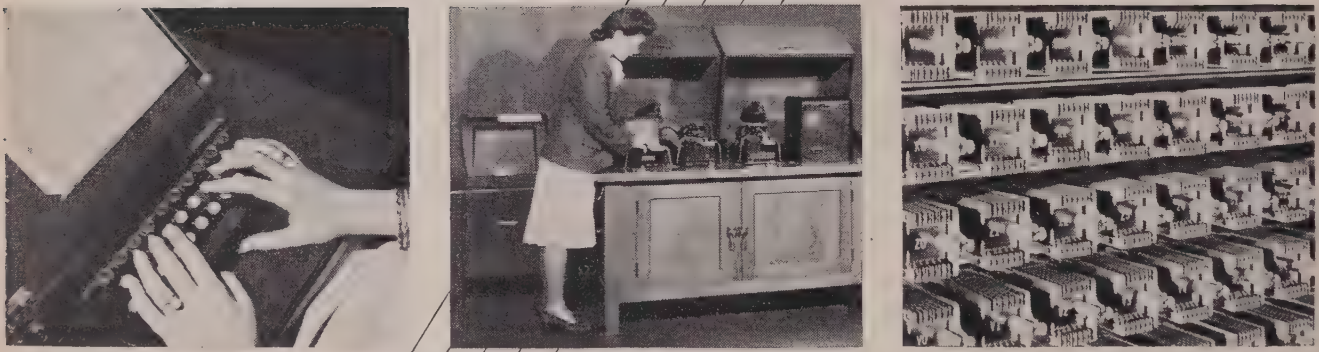
This exacting test is another good reason why Federal tubes are better tubes. Transmitting, rectifier, industrial power . . . they have a reputation that is deserved because they are *built to stay.*

Federal always *has* made better tubes.

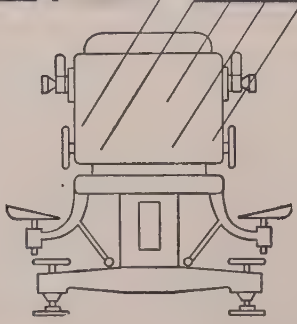
Federal Telephone and Radio Corporation

Newark 1, N. J.

Target practice with Relays and Keys



(Left to right) The operator punches the problem data on tape, which is fed into the computer. The solution emerges in the teletype receiver. Relays which figure out the problem look like your dial telephone system.



In designing the gun-control systems which shot down enemy planes, Army ballistic experts were faced by long hours of mathematical calculations.

So Bell Laboratories developed an electrical relay computer. It solved complicated problems more accurately and swiftly than 40 calculators working in shifts around the clock.

Resembling your dial telephone system, which seeks out and calls a telephone number, this brain-like machine selects and energizes electric circuits to correspond with the numbers fed in. Then it juggles the circuits through scores of combinations corresponding to the successive stages of long calculations. It will even solve triangles and consult mathematical tables. The operator hands it a series of problems with the tips of her fingers — next morning the correct answers are neatly typed. Ballistic experts used this calculator to compute the performance of experimental gun directors and thus to evaluate new designs.

In battle action, Electrical Gun Directors are, of course, instantaneous. Such a director helped to make the port of Antwerp available to our advancing troops by directing the guns which shot down more than 90% of the thousands of buzz bombs.

Every day, your Bell System telephone calls are speeded by calculators which use electric currents to do sums. Even now, lessons learned from the relay computer are being applied to the extension of dialing over toll lines.

BELL TELEPHONE LABORATORIES

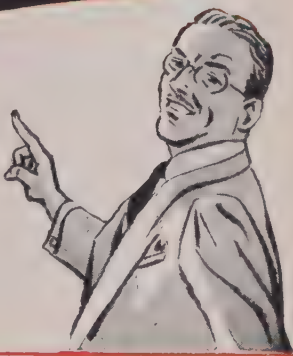
EXPLORING AND INVENTING, DEVISING AND PERFECTING FOR CONTINUED IMPROVEMENTS AND ECONOMIES IN TELEPHONE SERVICE

THE AMPEREXTRA FACTOR in SOUND TRANSMISSION

The Amperextra Factors of dependability and longevity represent important operational and replacement savings in the sound transmission field. Even in wartime, orders from essential civilian users were filled with fairly consistent regularity. Now, with nothing ahead but peace, the Amperextra Factor of service takes on an entirely new meaning for broadcasting stations, amateur radio operators and communications organizations. Your inquiries are invited.

WHAT ONE USER SAYS...

. . . "the ease with which they can be driven to full output, the simplification of cooling arrangements, the relative immunity to heavy overloads, and the moderate plate voltages required result in a combination not easily surpassed."



AMPEREX INTERCHANGEABILITY

Amperex tubes will fit into all types of transmitters for which they are intended, and may be interchanged or used to replace tubes of other manufacture without need for circuit readjustment and without impairment of transmitter performance.

SPECIALLY PROCESSED GRAPHITE ANODES..

. . . in many of our exclusive designs make for more uniform temperature distribution, absence of change in characteristics with time, and a higher initial vacuum which keeps tubes harder and assures longer life.

Amperex Type ZB-120 Transmitting Tube. Filament voltage, 10-10.5 volts AC or DC. Filament current, 2 amperes. Amplification factor, 90. Grid-to-Plate Transconductance at 120 ma., 5000 micromhos. Direct Interelectrode Capacitances: grid-to-plate, 5.2 μμf; grid-to-filament, 5.3 μμf; plate-to-filament, 3.2 μμf.

Amperex Type HF-3000 Transmitting tube. Filament voltage, 21 to 22. Filament current, 40.5 amperes. Filament emission, 6 amperes. Amplification factor, 16. Grid-to-Plate Transconductance of plate current of 1 ampere, 6500 micromhos. Direct Interelectrode Capacitances: grid-to-plate, 10 μμf; grid-to-filament, 13 μμf; plate-to-filament, 4 μμf.

Amperex Type 891-R Transmitting Tube. Filament, two-unit type for single-phase or two-phase AC or DC operation—voltage per unit, 11; current per unit, 60 amperes; amplification factor, 8. Grid-to-Plate Transconductance at a plate current of 0.75 ampere, 4000 micromhos. Direct Interelectrode Capacitances: grid-to-plate, 30 μμf; grid-to-filament, 16 μμf; plate-to-filament, 3 μμf.

AMPEREX

...THE HIGH PERFORMANCE TUBE

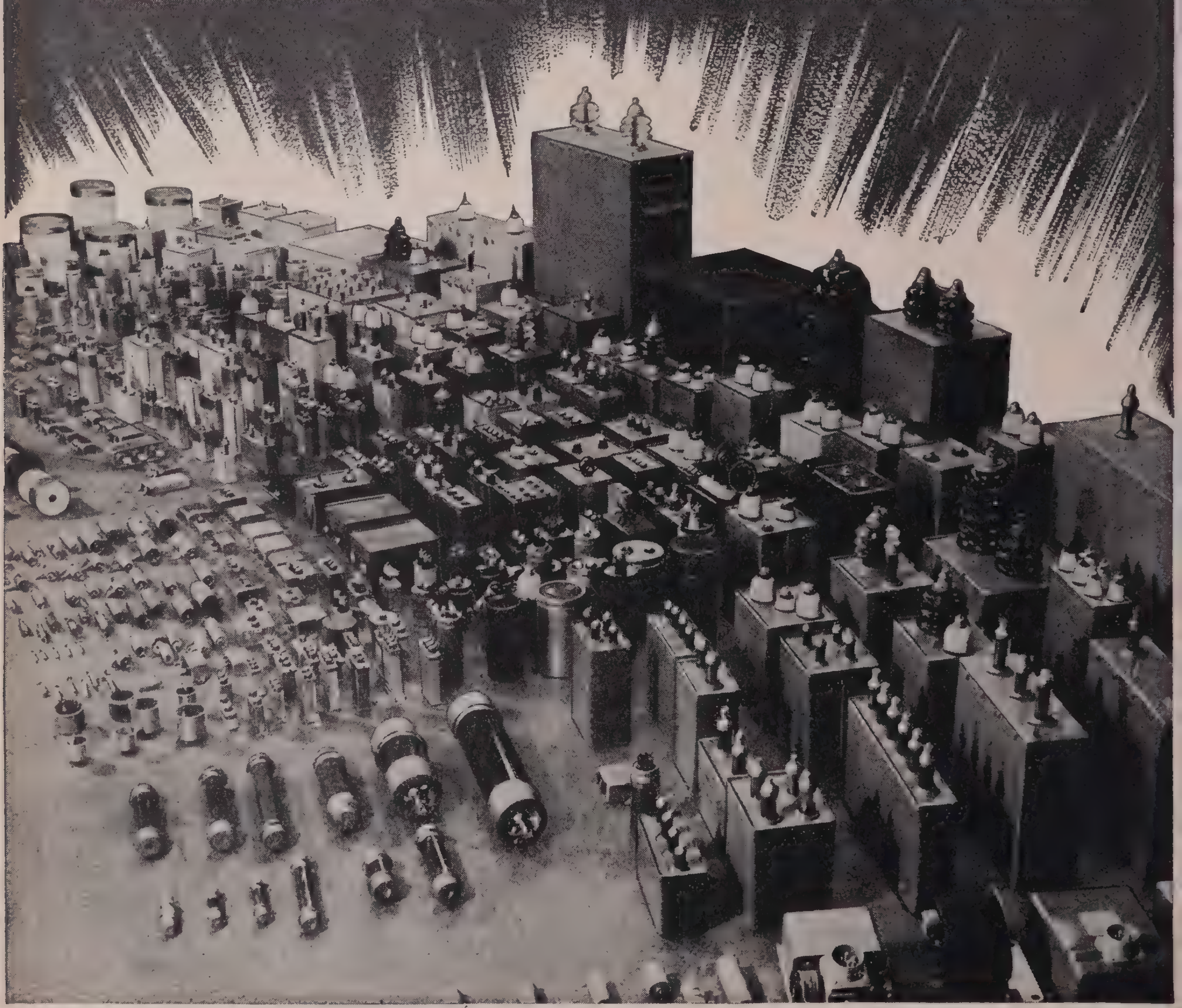
Many standard types of Amperex tubes are now available through leading radio equipment distributors. The Amperex Special Application Engineering Department will gladly work with you on the solution to your pressing problems.

AMPEREX ELECTRONIC CORPORATION

25 Washington St., Brooklyn 1, N. Y., Export Division: 13 E. 40th St., New York 16, N. Y., Cables: "Arlab"
Canadian Distributor: Rogers Majestic Ltd. • 622 Fleet Street West, Toronto

341 OF THE 9,675 CAPACITOR AND RESISTOR TYPES

engineered by SPRAGUE and produced in 1944



A good measure of any supplier is his ability to meet BOTH standard and highly specialized requirements. The Sprague wartime record offers convincing evidence in both respects.

CAPACITORS • *KOOLOHM RESISTORS • *CEROC 200 INSULATION

ARMY E NAVY

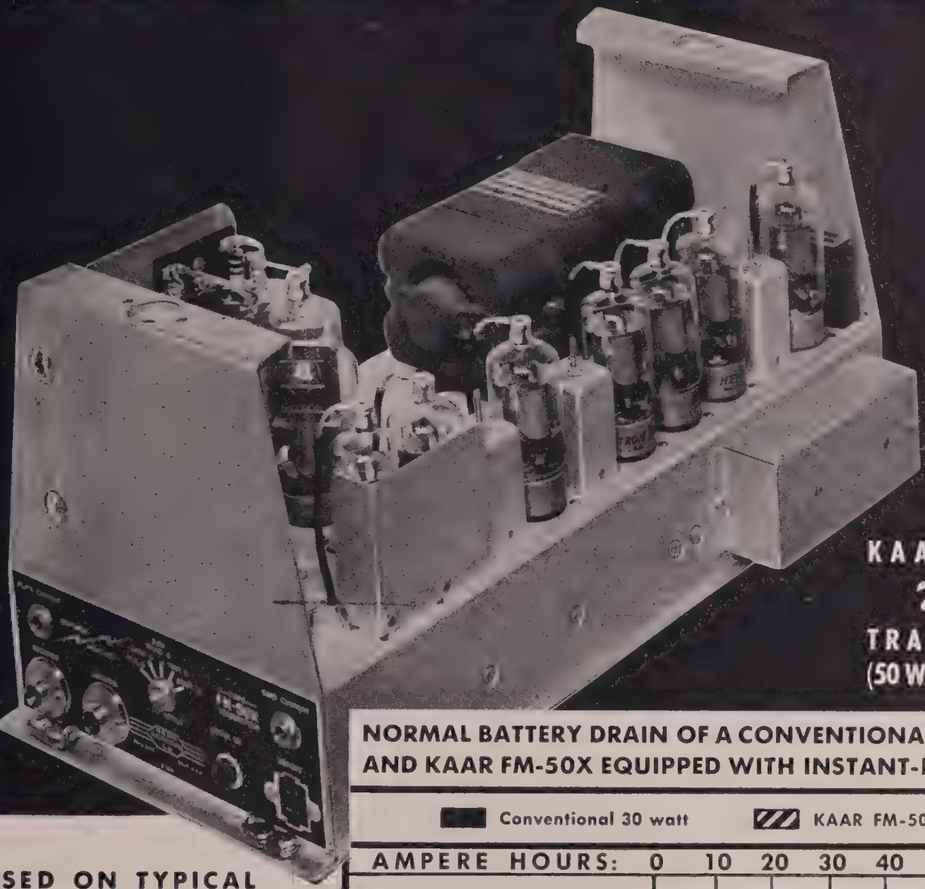
SPRAGUE

SPRAGUE
ELECTRIC COMPANY
NORTH ADAMS, MASS.

**Trademarks Registered U. S. Patent Office*

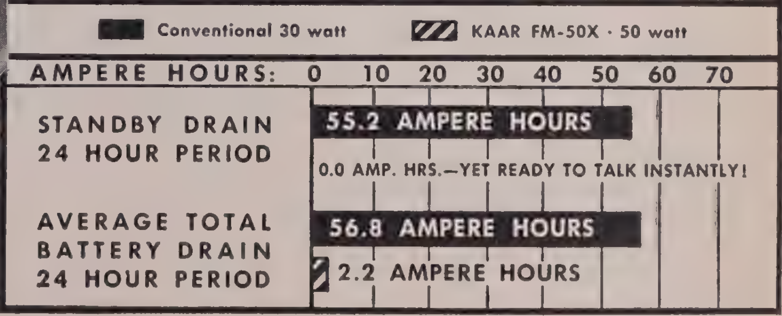
PIONEERS OF ELECTRIC & ELECTRONIC PROGRESS

Compare the actual battery drain!*



KAAR FM-50X
Mobile
TRANSMITTER
(50 WATTS OUTPUT)

NORMAL BATTERY DRAIN OF A CONVENTIONAL TRANSMITTER AND KAAR FM-50X EQUIPPED WITH INSTANT-HEATING TUBES



*** CHART BASED ON TYPICAL METROPOLITAN POLICE USE**

(140 Radiotelephone-equipped cars operating three shifts in city of 600,000 population.)

MESSAGES ORIGINATED BY CARS	904
MESSAGES ACKNOWLEDGED BY CARS	932
TOTAL TRANSMISSIONS PER CAR	13
AVE. LENGTH OF TRANSMISSION	15 sec.
AVE. TRANSMITTING TIME 24 HOURS	3 min. 15 sec.

KAAR mobile FM-50X transmitter gives you 20 watts more output with only 1/25th usual battery drain!

KAAR engineers—who pioneered the instant-heating AM radiotelephone—have now, through the use of instant-heating tubes, made 50 and 100 watt *mobile* FM transmitters practical! Thus you gain greater power and range—along with a tremendous reduction in battery drain!

With instant-heating KAAR equipment standby-current is zero—yet the moment you press the button microphone you are on the air. Contrast this with conventional emergency transmitters, over 90% of which operate with the filaments "hot" during stand-by. Since sturdy instant-heating tubes eliminate this great waste of energy without slowing the handling of messages, KAAR 50 and 100 watt transmitters can be operated from the standard ignition battery!

100 WATT MOBILE FM!

The KAAR FM-100X is identical to the FM-50X, except for the final amplifier. It puts 100 watts into a standard 34 ohm non-inductive load and is ideal for county and state police use. It requires no special batteries, wiring, or generator changes.

ADDITIONAL FEATURES

A new system of modulating the phase modulator tubes in KAAR FM transmitters provides excellent voice quality. Note that the equipment is highly accessible, and only two types of tubes are used. Frequency range: 30 to 44 megacycles.

Write today for free bulletin describing KAAR FM transmitters in detail. It's ready now!

KAAR ENGINEERING CO.

PALO ALTO CALIFORNIA

Export Agents: FRAZAR AND HANSEN · 301 Clay St · San Francisco, Calif.

NEW RAYTHEON SUB-MINIATURE TUBES for Pocket Receivers

RAYTHEON
2E32
5AX2
ACTUAL SIZE
RAYTHEON
2G22
7AX2
ACTUAL SIZE
RAYTHEON
2E42
6AX2
ACTUAL SIZE
RAYTHEON
2E36
5AX2
ACTUAL SIZE

Since 1938 Raytheon has pioneered in the design, development and production of small, low drain, long life tubes. These have helped to make possible the modern extremely compact hearing aid.

Now for Radio Receivers—Now Raytheon announces a physically similar kit of flat style, sub-miniature tubes for radio receiver applications. Included is a shielded RF-pentode amplifier, a triode-heptode converter, a diode-pentode detector-amplifier and an output pentode for earphone operation.

Much Smaller Radios Possible—These tubes make it possible to construct radios a fraction the size of prewar "personals," with sensitivity rivaling much larger sets.

The ratio of performance to battery drain is maintained very high, thus assuring the maximum possible operating life from the small size batteries now available.

The line consists of tubes approximately 1$\frac{9}{16}$″ long x 0.3″ x 0.4″ in cross section. Each type is available with pins for use with small commercially available sockets as illustrated, or may be had with long flexible leads for wiring the tube directly into the circuit.

No progressive radio manufacturer will overlook the tremendous possibilities inherent in the small pocket receiver—built around the new Raytheon sub-miniature tubes. But call on Raytheon for *every* tube need—large or small—for the finest in engineering, production and performance.

ELECTRICAL CHARACTERISTICS

	2E31† 2E32# Shielded RF Pentode	2G21† 2G22# Triode-Heptode	2E41† 2E42# Diode-Pentode	2E35† 2E36# Output Pentode
Filament Voltage	1.25 V	1.25 V	1.25 V	1.25 V
Filament Current	50 ma	50 ma	30 ma	30 ma
Max. Grid-Plate Capacitance	0.018 µµf	0.065 µµf‡	0.10 µµf	0.2 µµf
Plate Voltage**	22.5 V	22.5 V	22.5 V	22.5 V
Screen Voltage	22.5 V	22.5 V	22.5 V	22.5 V
Control Grid Voltage*	0	0	0	0
Osc. Plate Voltage	—	22.5 V	—	—
Plate Current	0.35 ma	0.2 ma	0.4 ma	0.27 ma
Screen Current	0.3 ma	0.3 ma	0.15 ma	0.07 ma
Osc. Plate Current	—	1.0 ma	—	—
Transconductance	500 µmhos	60 µmhos (Gc)	400 µmhos	385 µmhos
Plate Resistance	0.35 meg	0.5 meg##	0.25 meg	0.22 meg

*With 5 megohm grid resistance connected to F−.
**Higher voltage operation is possible as shown on engineering characteristics sheet available by request.

†Flexible lead Types.
#Plug-in Types.
##Approximate conversion Rp.
‡Signal grid to mixer plate Capacitance

RAYTHEON
MANUFACTURING COMPANY

Radio Receiving Tube Division

NEWTON, MASSACHUSETTS • LOS ANGELES
NEW YORK • CHICAGO • ATLANTA

EXCELLENCE IN ELECTRONICS

THIS NEW MIGHTY MIDGET

has changed the lives of this whole family of "MINI-MAX" BATTERIES

No. 412
MINI-MAX
"B" BATTERY
FOR HEARING AIDS
MINIATURE RADIOS, ETC.
No. 482
45 VOLTS
No. 482
EVEREADY
MINI-MAX
"B" BATTERY
FOR
PORTABLE
RADIOS
No. 467
67½ VOLTS
No. 467
EVEREADY
MINI-MAX
"B" BATTERY
FOR MINIATURE RADIOS
...HEARING AIDS, ETC...
No. 420
22½ VOLTS
MINI-MAX
"B" BATTERY
FOR HEARING AIDS
MINIATURE RADIOS, ETC.

The four batteries shown above are approximately ⅔ of actual size

EVEN BEFORE PEARL HARBOR, battery construction principles developed by National Carbon Company were making possible new strides in portable radio and electronic equipment. Then came the war. The company was called upon to develop even more radical improvements in battery construction to meet the needs of light and extremely portable military communications of all types, and so the tiny 22½ volt "Eveready" "Mini-Max" "B" battery was born—a battery well under half the size of anything of comparable voltage—easy to carry as a match box!

This is what this new, improved battery construction means. It means a brand new line of portable radio equipment—equipment that will give the idea of the "personal radio" an entirely new meaning. It makes possible radio sets for individual use—sets so small that they can be slipped into the pocket of a vest, or carried in a woman's handbag. Portable radio business will not merely pick up where it left off December 7, 1941. It will be years ahead of itself.

Engineers and designers are already aware of the possibilities of this new battery. They are already at work on new radio and electronic devices which exploit its portability. And at this time may we invite all these creative men to avail themselves of our experience, our laboratories and to consult with our engineers. National Carbon Company, Inc. extends to you complete cooperation.

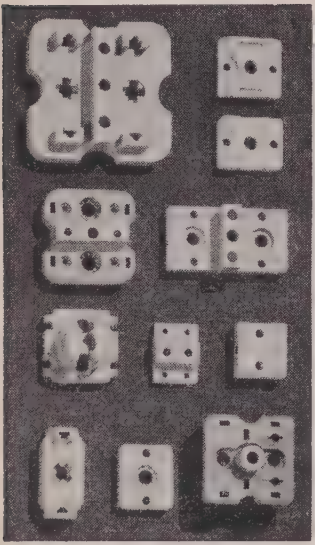
The words "Eveready" and "Mini-Max" are registered trade-marks of National Carbon Company, Inc.

Now that radio batteries are back again, National Carbon Company is offering an extremely useful new Portable Radio Battery Replacement Guide. Write for your copy today to our nearest Division Office listed below.

NATIONAL CARBON COMPANY, INC.
Unit of Union Carbide and Carbon Corporation

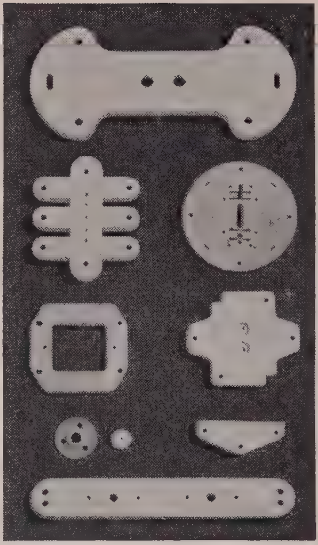
General Offices: NEW YORK, N. Y.
Division Sales Offices: Atlanta, Chicago, Dallas, Kansas City, New York, Pittsburgh, San Francisco.

STUPAKOFF
CERAMIC INSULATORS FOR YOUR PRODUCTS



PADDER AND TRIMMER BASES

Padder and trimmer bases that are mechanically strong, dimensionally accurate, and electrically stable, keep assembly lines flowing, minimize breakage in production and in use, insure consumer satisfaction. STUPAKOFF combines mass production with laboratory precision. Exacting control extends from scientific testing of raw material through final packing. STUPAKOFF engineers are at your service.

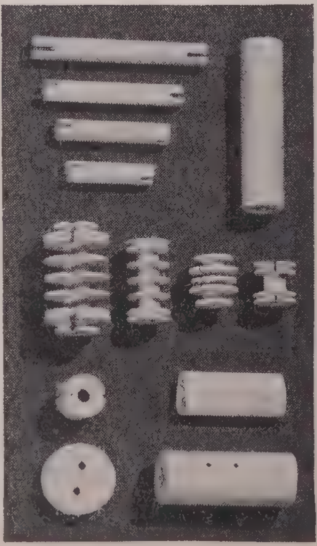


TUBE PARTS

From the day when the first STUPAKOFF ceramic heater insulator was produced for the first A.C. radio tube, the name STUPAKOFF has been a synonym for quality in the field of radio ceramics. Adherence to specification tolerances, both mechanical and electrical, and to the proper material for the specific application are integrated in every STUPAKOFF product.

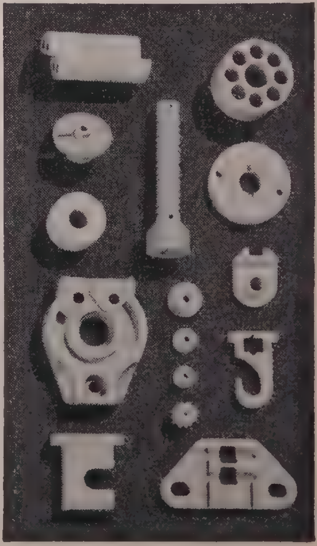
RESISTOR PARTS

STUPAKOFF produces precision-made ceramic resistor parts as rods, plain or threaded; as tubes, plain or threaded; as winding forms for *all* types of resistors; and metallized for solder-sealed resistors. STUPAKOFF ceramics are dense and sturdy, vitrified to withstand moisture, resistant to vibration and thermal shocks.



ELECTRIC APPLIANCE CERAMICS

STUPAKOFF insulators are planned to meet the demands of assembly line production and to endure the rigorous usage requirements, thus minimizing field failures and service calls. Made vitrified, dense, non-hygroscopic or porous as required. Made to withstand sudden temperature changes without fracturing. Engineered to suit the job.

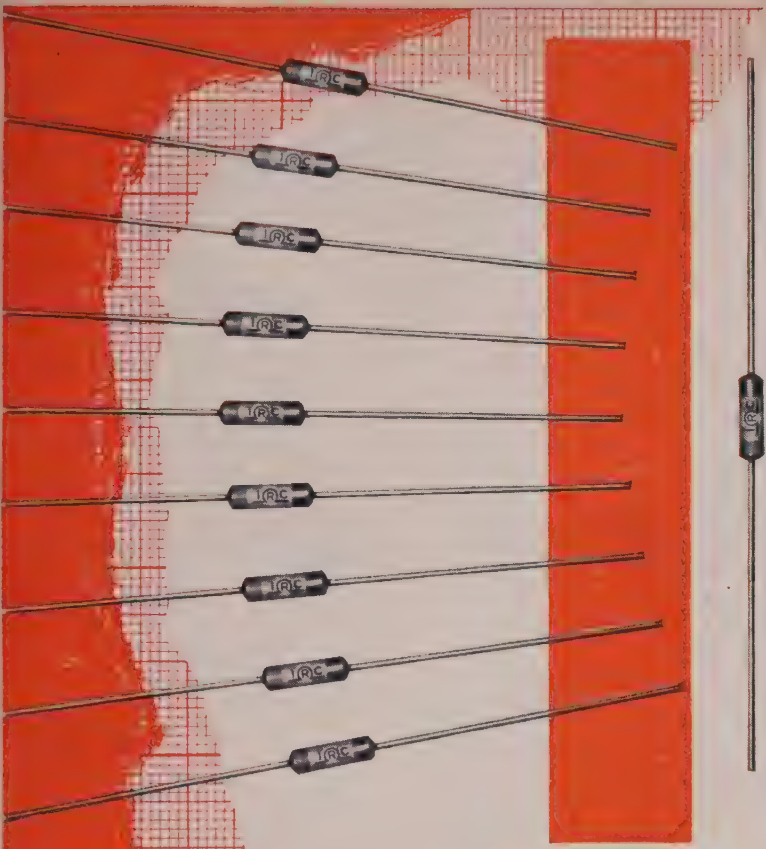


★ BUY VICTORY BONDS ★

"FOR GREAT ACHIEVEMENT"

STUPAKOFF CERAMIC AND MANUFACTURING CO., LATROBE, PA.

Products for the World of Electronics



NOW
IT CAN BE
S
TOLD

IRC TYPE BWS
¼ WATT INSULATED WIRE WOUND RESISTORS

Here's a brand new IRC resistor that until a few weeks ago was very "hush-hush" except to a few selected laboratories and prime contractors engaged in development and manufacture of VT proximity fuzes. Production on this small, efficient unit, in the last several months before V-J day, mounted to amazing figures to keep pace with the advancing victory tempo. Now this same high quality resistor is available in quantity to help solve many a "small space" resistance problem for you. Rated a full quarter watt, small 'round as a match stick and but 13/32" in length, this sturdy insulated wire wound can be depended upon for "Preferred Performance." For technical data refer to IRC Engineering Bulletin No.3, available on request from *Dept. 10-L.*

INTERNATIONAL RESISTANCE CO.
401 NORTH BROAD STREET • PHILADELPHIA 8, PA.

In Canada: International Resistance Co., Ltd., Toronto

PREFERRED FOR PERFORMANCE • IRC • FIXED & VARIABLE RESISTORS •

IRC makes more types of resistance units, in more shapes, for more applications, than any other manufacturer in the world.

CAPACITOR *Craftsmanship*

each
the specialized product of specialists... yet available from ONE dependable source of supply

● With tiny silvered micas, it's precision: capacitance tolerances of 1%, with temperature coefficients and stability requirements to meet the highest characteristics requirements of JAN-C-S; excellent retrace characteristics; practically no capacitance drift with time; exceptionally high Q. Yes, Aerovox *specializes* in such precision capacitors.

And at the other extreme are giant Type 26 oil-filled capacitors for high-voltage requirements such as in X-ray equipment, high-voltage test and laboratory equipment, and for carrier-current coupling. Again, Aerovox *specializes* in high-voltage oil-impregnated, oil-filled capacitors.

But how, you ask, can one organization really *specialize* in such totally different products? The Aerovox answer:

The huge Aerovox plant is really *several* plants in one. Micas are made in the Mica Department, oils in the Oil Department, electrolytics in the Electrolytic Department, and so on. Each has its OWN engineers, supervisors, trained workers.

Thus you are assured of that *specialized* craftsmanship that insures the best in highly specialized products, along with the convenience, certainty and economy of ONE outstanding source of supply.

● ***Try us on that capacitance problem***

AEROVOX *Capacitors*

INDIVIDUALLY TESTED

AEROVOX CORPORATION, NEW BEDFORD, MASS., U. S. A. • SALES OFFICES IN ALL PRINCIPAL CITIES

Export: 13 E. 40 St., New York 16, N. Y. • Cable: 'ARLAB' • In Canada: AEROVOX CANADA LTD., HAMILTON, ONT.

an ENTIRELY NEW concept
in CERAMIC TRIMMER designs
by ERIE RESISTOR

ERIE STYLE TS1F CERAMIC TRIMMER

HERE is a new ceramic trimmer that's unique in design, extremely compact, with desirable capacity ratios and is priced for a wide range of applications in broadcasting and high frequency bands.

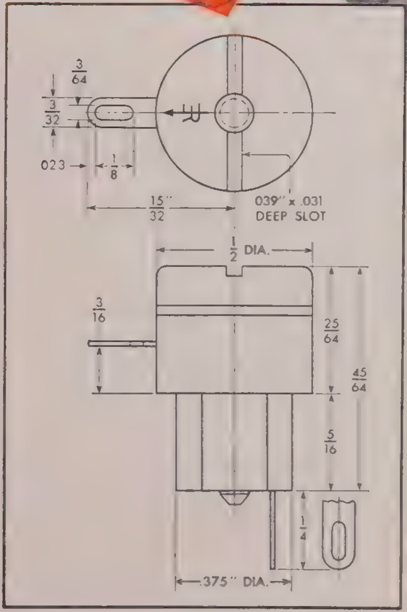
The Erie TS1F trimmer employs a ceramic dielectric and is available in nominal temperature coefficients, zero, −300, and −750 parts/million/°C. In the N750 coefficient, capacity range of 8-50 MMF is available. Corresponding ratios in lower capacities are furnished with zero and N300 coefficient.

Capacity change is essentially constant per degree of rotation, and full range is covered in 180° rotation.

The metal rotor completely covers the stator track. Contact surfaces of both rotor and stator are lapped, providing a high degree of stability, preventing dust or other foreign matter from affecting the performance characteristics of the unit, and keeping noise level to a minimum at high frequencies. Electrical characteristics are given at the right.

These trimmers are firmly held in place in a D hole in the chassis by means of a multiple-tooth spring clip, furnished with the trimmer.

For complete information contact our nearest representative or write us direct.



ERIE TS1F CERAMICON TRIMMER

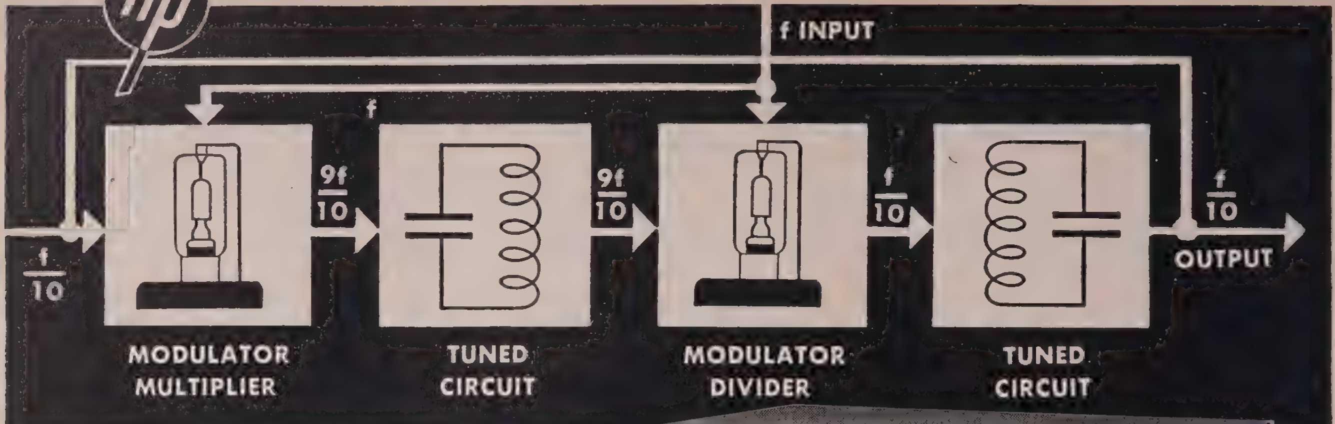
Voltage Rating: 350 volts D.C.
Flash Test: 700 volts D.C. for 15 seconds
Initial Q Factor at 1MC: 500 minimum
Initial Leakage Resistance: 10,000 meg. min.

Electronics Division

ERIE RESISTOR CORP., ERIE, PA.

LONDON, ENGLAND   TORONTO, CANADA.

LABORATORY INSTRUMENTS FOR SPEED AND ACCURACY

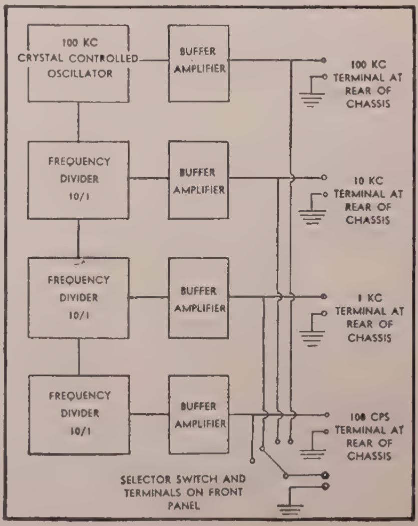

f INPUT
f
$\frac{f}{10}$
$\frac{9f}{10}$
$\frac{9f}{10}$
$\frac{f}{10}$
$\frac{f}{10}$
OUTPUT
MODULATOR MULTIPLIER
TUNED CIRCUIT
MODULATOR DIVIDER
TUNED CIRCUIT

THIS CIRCUIT PROVIDES AN ACCURATE FIXED STANDARD FREQUENCY . . .

A modulator divider tube with a resonant circuit tuned to 1f/10 and a modulator multiplier tube with a resonant circuit of 9f/10 are the fundamentals of a frequency divider unit which is the basic element of this *-hp-* Secondary Frequency Standard.

A small transient voltage in the resonant circuit of the modulator divider tube is applied to the grid of the modulator multiplier tube, and the input voltage f is also applied to this tube. The two voltages mix to supply an output frequency of 9f/10. This frequency (9f/10) is fed to the grid of the modulator divider tube where it is mixed with the input frequency (f), and results in a frequency of 1f/10 in the modulator divider tuned circuit. The action is repeated and the voltage is built up until a stabilized condition is reached or until the frequency (f) is removed. Thus the output of the divider unit is controlled by the input frequency.

Three such frequency divider circuit units in conjunction with a temperature controlled oscillating quartz crystal, which generates 100 kc, make up the *-hp-* 100B Secondary Frequency Standard. By cascading the 100 kc down through the three dividers, accurate fixed frequencies of 10 kc, 1 kc and 100 cps are made available in addition to the 100 kc supplied by the oscillator.


100 KC CRYSTAL CONTROLLED OSCILLATOR
BUFFER AMPLIFIER
100 KC TERMINAL AT REAR OF CHASSIS
FREQUENCY DIVIDER 10/1
BUFFER AMPLIFIER
10 KC TERMINAL AT REAR OF CHASSIS
FREQUENCY DIVIDER 10/1
BUFFER AMPLIFIER
1 KC TERMINAL AT REAR OF CHASSIS
FREQUENCY DIVIDER 10/1
BUFFER AMPLIFIER
100 CPS TERMINAL AT REAR OF CHASSIS
SELECTOR SWITCH AND TERMINALS ON FRONT PANEL

As can be noted by the block diagram, these frequencies are available through a selector switch (on front of panel) or individually from binding posts (rear of chassis). All four fixed frequencies can be utilized at separate test stations simultaneously, which is an economical feature. This instrument is extremely valuable for use in audio and the low radio frequency fields. More complete information will be gladly sent in response to your inquiry. 1072

These -hp- Representatives are at your Service

Eastern Representative:
Burlingame Associates, Ltd.
11 Park Place
New York 7, N. Y.—Worth 2-5563-4

Midwestern Representative
Alfred Crossley
549 West Randolph St.
Chicago 6, Ill.—State 7444

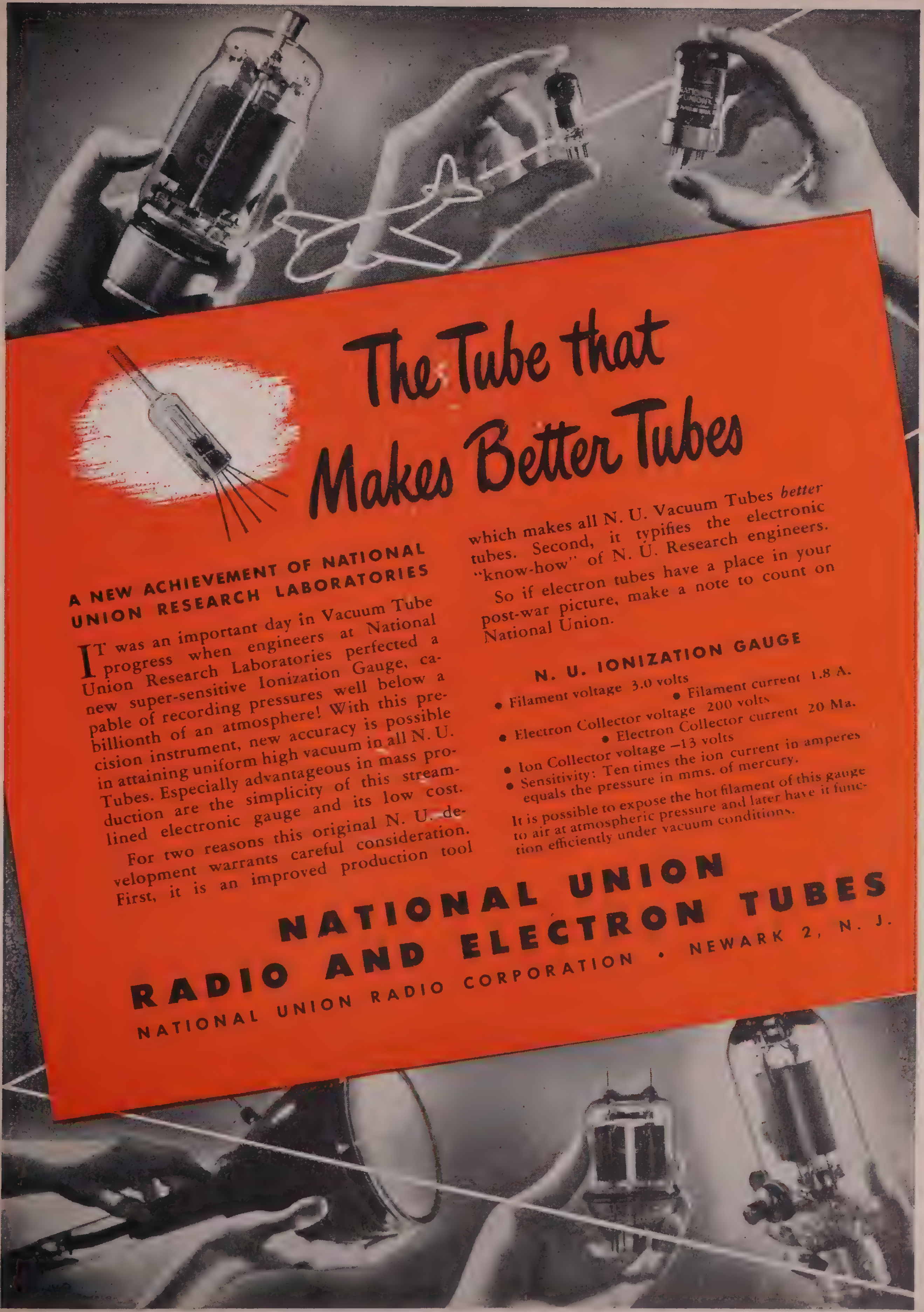
Western Representative:
Norman B. Neely Enterprises
7422 Melrose Ave.
Hollywood 46, Calif.—Hillside 9133

Canadian Representative:
Atlas Radio Corporation
560 King Street West
Toronto 2, Canada—Wa 4761

HEWLETT-PACKARD COMPANY

BOX 1072E · STATION A · PALO ALTO, CALIFORNIA

Audio Frequency Oscillators *Signal Generators* *Vacuum Tube Voltmeters*
Noise and Distortion Analyzers *Wave Analyzers* *Frequency Meters*
Square Wave Generators *Frequency Standards* *Attenuators* *Electronic Tachometers*



The Tube that Makes Better Tubes

A NEW ACHIEVEMENT OF NATIONAL UNION RESEARCH LABORATORIES

IT was an important day in Vacuum Tube progress when engineers at National Union Research Laboratories perfected a new super-sensitive Ionization Gauge, capable of recording pressures well below a billionth of an atmosphere! With this precision instrument, new accuracy is possible in attaining uniform high vacuum in all N. U. Tubes. Especially advantageous in mass production are the simplicity of this streamlined electronic gauge and its low cost.

For two reasons this original N. U. development warrants careful consideration. First, it is an improved production tool which makes all N. U. Vacuum Tubes *better* tubes. Second, it typifies the electronic "know-how" of N. U. Research engineers.

So if electron tubes have a place in your post-war picture, make a note to count on National Union.

N. U. IONIZATION GAUGE

- Filament voltage 3.0 volts
- Filament current 1.8 A.
- Electron Collector voltage 200 volts
- Electron Collector current 20 Ma.
- Ion Collector voltage −13 volts
- Sensitivity: Ten times the ion current in amperes equals the pressure in mms. of mercury.

It is possible to expose the hot filament of this gauge to air at atmospheric pressure and later have it function efficiently under vacuum conditions.

NATIONAL UNION RADIO AND ELECTRON TUBES

NATIONAL UNION RADIO CORPORATION • NEWARK 2, N. J.

1895
1896
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THE
INSIDE STORY

A background of *Performance*—over 50 years—is the *inside story* of the popularity that has brought leadership to *Thordarson* transformers. Performance over the years, after all, is the only true test of product quality.

Consumer acceptance will continue because *Thordarson* research and design engineers are never satisfied just keeping abreast of the times. These men are continually developing many transformer components which are instrumental in the production of new and better performing devices and equipment for the electronics industry.

This same pioneering spirit has been responsible for many new *Thordarson* transformer applications and developments during the war • • • all of which will be available shortly for civilian requirements.

Thordarson's well-tested methods of sales promotion and distribution will continue their joint task of making *Thordarson* Transformers, together with complete information on their applications and use, available to everyone in the field.

Always think of Thordarson for top-notch transformers!

500 WEST HURON ST., CHICAGO, ILL.

T

ARMY E NAVY

ORIGINATORS OF TRU-FIDELITY AMPLIFIERS

THORDARSON

ELECTRIC MANUFACTURING DIVISION

MAGUIRE INDUSTRIES, INCORPORATED

MORE POWER OUTPUT *BUT* LESS BATTERY DRAIN WITH HYTRON INSTANT-HEATING BEAM TETRODES

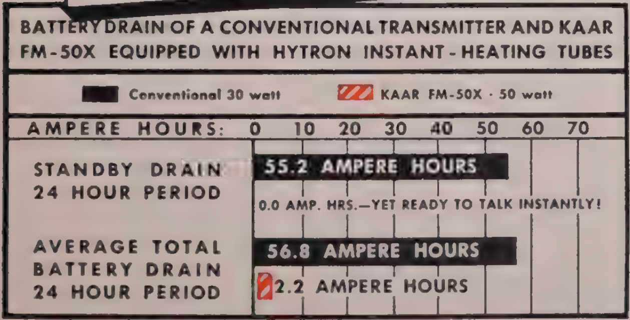
ZERO STAND-BY CURRENT Thoriated tungsten filaments of the Hytron 2E25, HY69, and HY1269 permit simultaneous application of all potentials. During stand-by, no precious filament current is drawn from the battery. Especially with the larger tube complements of FM transmitters, is conservation of battery power mandatory.

MORE OUTPUT—GREATER RANGE Only 4% of the current required for cathode types, is necessary to operate the instant-heating 2E25, HY69, and HY1269. (See table below.) Even in a mobile FM transmitter, 100 watts output is practicable. Imagine the advantages of such increased output in police, marine, or other mobile equipment.

SPARES PROBLEM SIMPLIFIED Using the 2E25, HY69, and HY1269, you take full advantage of the beam tetrode's versatility. The 2E25, for example, can power a whole transmitter—AF and RF—AM or FM. If more output is required, HY69's or HY1269's in push-pull still confine the spares complement to only two types.

ADVANTAGES OVER CATHODE TYPES Yes, the 2E25, HY69, and HY1269 cost more than cathode types. But they are worth it. Not only are they easier on the battery, and permit larger outputs, but they are designed, built, and tested for transmitting. Some advantages are: centering of filament potential at 6.0 volts, r.f. shielding to eliminate the necessity for neutralization, low-loss insulation throughout, plate connection to top cap, and rugged construction.

HY69
HY 1269
2E25
HYTRON
MADE IN U.S.A.
$3.95
$4.50
$3.95



This chart, prepared by Kaar Engineering Co., is based on typical metropolitan police use of 140 radiotelephone-equipped cars operating three shifts in a city of 600,000 population. The 24-hour survey included 904 messages originated by cars and 932 messages acknowledged by cars. Transmissions averaged: 13 per car, 15 seconds in length, and 3 minutes 15 seconds transmitting time.

ABBREVIATED DATA
HYTRON INSTANT-HEATING BEAM TETRODES

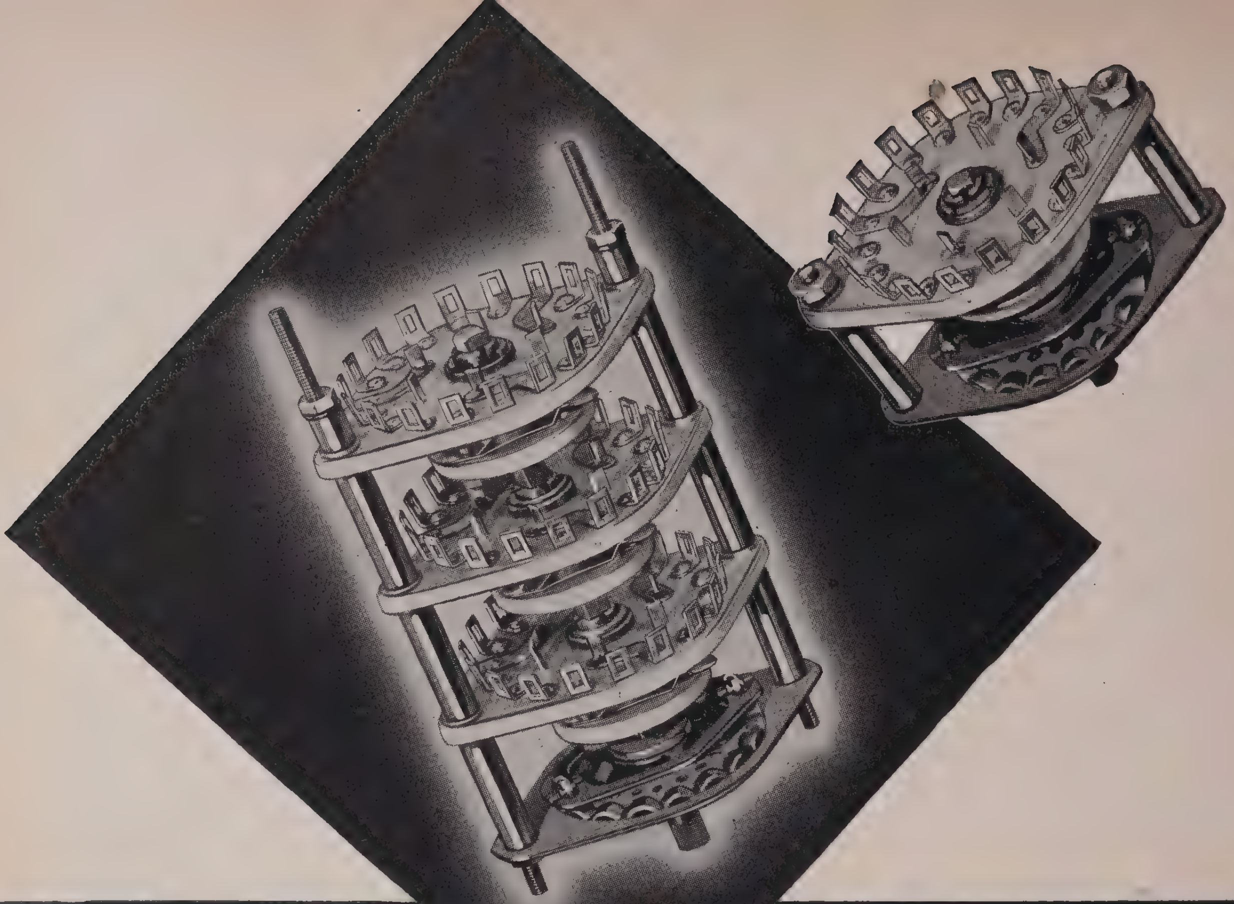
Characteristic	*2E25*	*HY69*	*HY1269*
Filament Potential (volts)	6.0	6.0	6/12
Filament Current (amps.)	0.8	1.6	3.2/1.6
Plate Potential (max. volts)	450	600	750
Plate Current (max. ma.)	75	100	120
Plate Dissipation (max. watts)	15	30	30
Grid-to-Plate Capacitance (mmfd.)	0.15	0.25	0.25
Maximum Seated Height (inches)	3 5/8	5 1/4	5 1/4
Maximum Diameter (inches)	1 7/16	2 1/16	2 1/16
Class C Power Output (watts)	24	42	63
Class C Driving Power (watts)	Less than one watt		

OLDEST MANUFACTURER SPECIALIZING IN RADIO RECEIVING TUBES

HYTRON

RADIO AND ELECTRONICS CORP.

MAIN OFFICE: SALEM, MASSACHUSETTS



Centralab

CRL

MEDIUM DUTY POWER SWITCHES

. . . available for transmitters, power supply converters and many special industrial and electronic uses.

The units are assembled in multiple gangs with a choice of shorting or non-shorting contacts.

The switching combinations manufactured for stock delivery are in single or multiple sections . . . 3 pole, 5 positions and 1 pole, 17 positions. (17 positions can be furnished with 18 positions continuous rotation.) Special combinations available.

Rated at 7½ amperes at 60 cycles, 115 volts, voltage breakdown 2500 volts D.C. to ground.

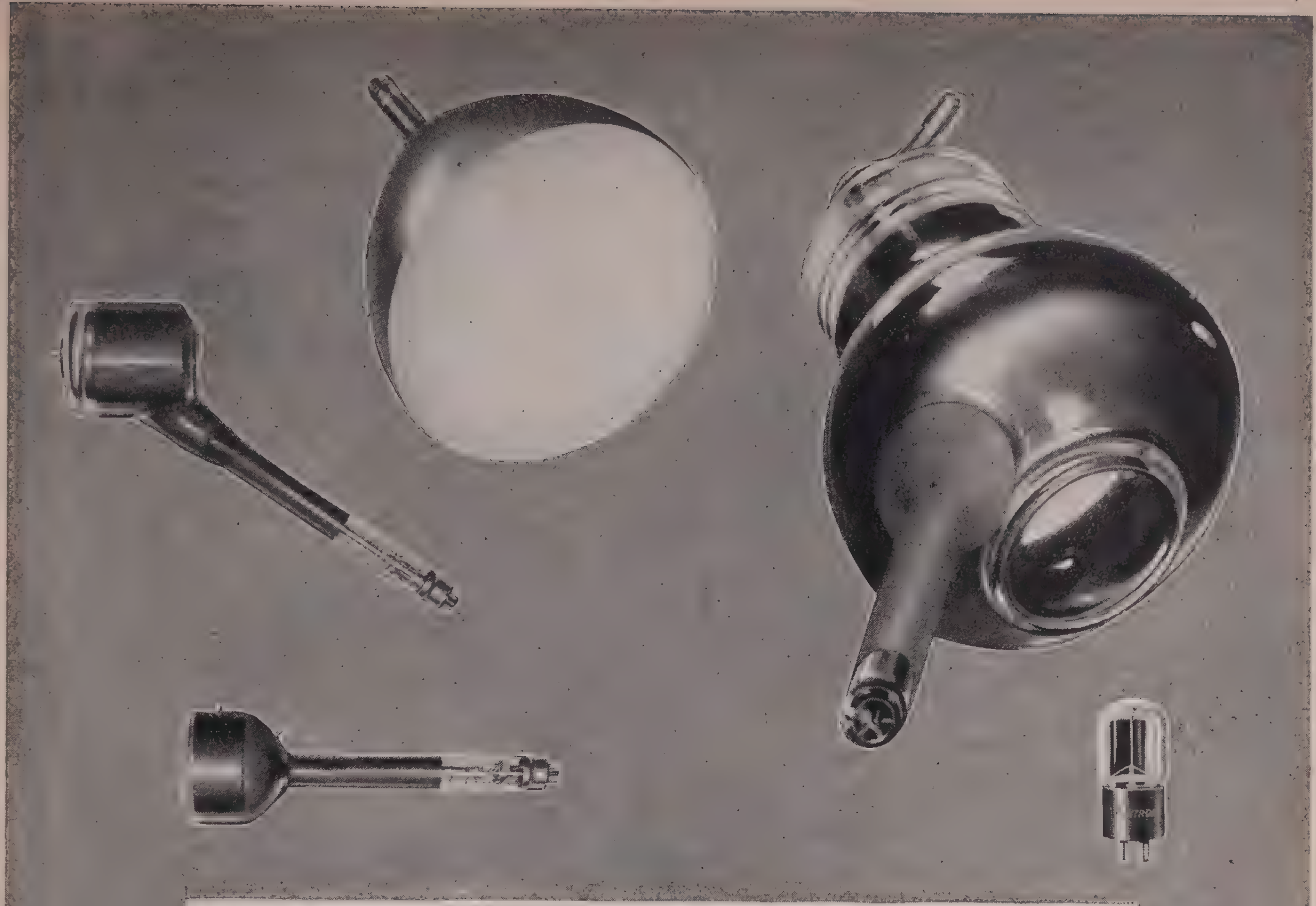
Write for Bulletin 815.

Producers of:
Variable Resistors •
Selector Switches •
Ceramic Capacitors,
Fixed and Variable
Steatite Insulators
and Silver Mica Button-type Capacitors.

Centralab

CRL

Division of GLOBE-UNION INC., Milwaukee



VISITRON

Television Tubes

by

Visitron is not a new name in tubes. *Visitron* is Rauland's name for all electronic tubes made in the Rauland Tube Division. It is the mark of the advanced *Rauland* Television thinking and planning based upon a pioneering experience second to none. *Rauland Visitron* tubes for direct-viewing for the home and projection for the home and theatre are ready to take their places in the new era of Television entertainment now unfolding before us. To be sure of your tube, be sure it's "*Visitron*."

RADIO · RADAR · SOUND Rauland **COMMUNICATIONS · TELEVISION**

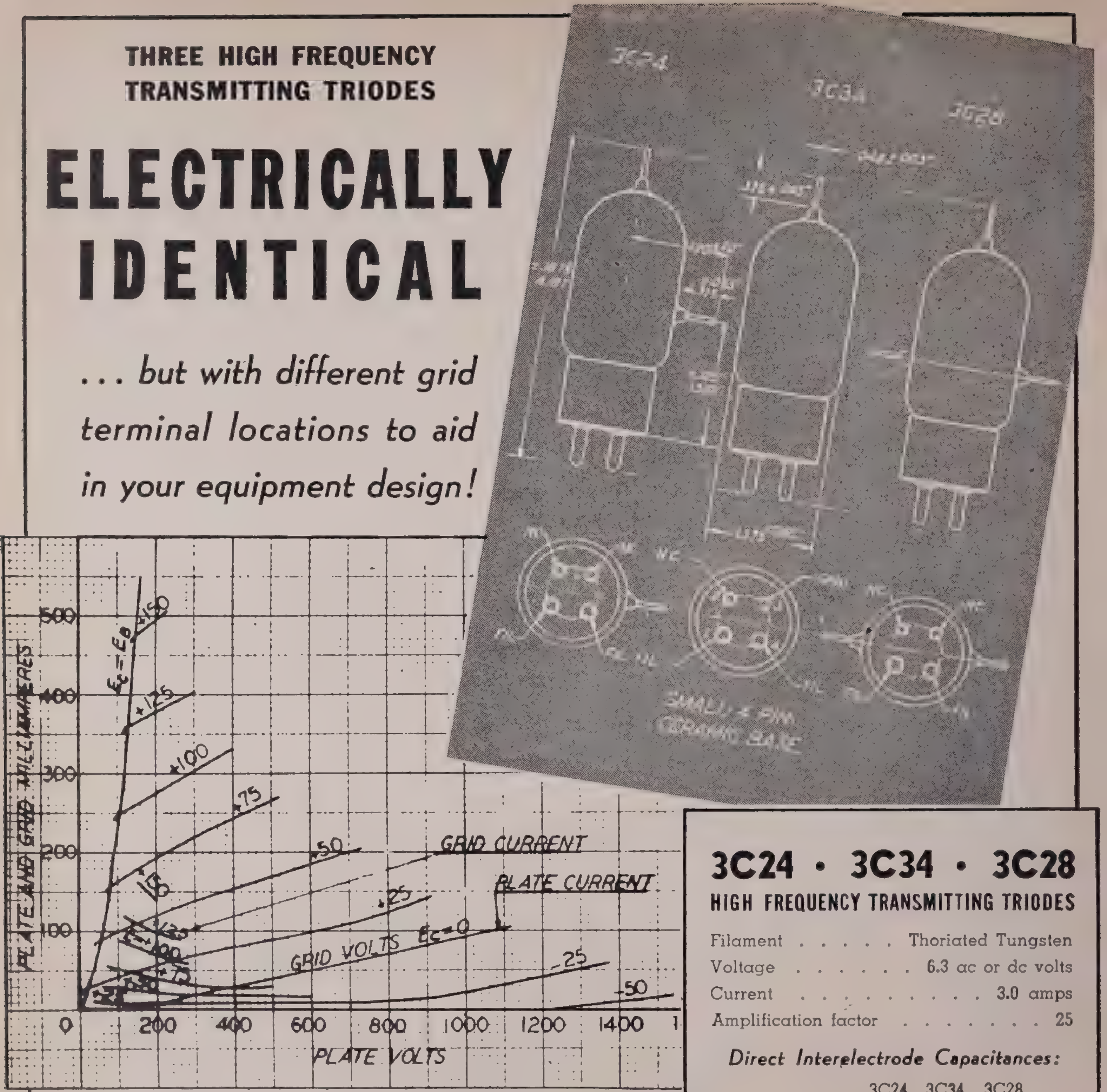
Electroneering is our business

THE RAULAND CORPORATION • CHICAGO 41, ILLINOIS

THREE HIGH FREQUENCY TRANSMITTING TRIODES

ELECTRICALLY IDENTICAL

... but with different grid terminal locations to aid in your equipment design!


3C24
3C34
3C28
SMALL 4 PIN CERAMIC BASE
$E_c=E_B$
+150
+125
+100
+75
+50
+25
GRID CURRENT
PLATE CURRENT
GRID VOLTS
$E_c=0$
-25
-50
150
125
100
75
500
400
300
200
100
PLATE AND GRID MILLIAMPERES
0
200
400
600
800
1000
1200
1400
PLATE VOLTS

• Here, again, Lewis engineers offer wider latitude in equipment design through the development of these three electrically identical high frequency transmitting triodes . . . with different grid terminal locations. If you are interested in these, or other type tubes with special design features applicable to your equipment, write, wire or phone today.

3C24 • 3C34 • 3C28

HIGH FREQUENCY TRANSMITTING TRIODES

Filament	Thoriated Tungsten
Voltage	6.3 ac or dc volts
Current	3.0 amps
Amplification factor	25

Direct Interelectrode Capacitances:

	3C24	3C34	3C28	
Grid-Plate	1.6	1.7	1.8	mmfds
Grid-Filament	2.0	2.5	2.1	mmfds
Plate-Filament	0.2	0.4	0.1	mmfds

Class "C" R-F Amplifier Frequency Limits

	3C24	3C34	3C28	
Full power input	60	60	100	mcs
Half power input	250	200	350	mcs

Licensed under R.C.A. Patents

Lewis ★ ELECTRONICS
LOS GATOS · CALIFORNIA

NOW READY—

RCA
VOLUME
SQUELCH
RCA
OFF

ABOVE: Station Handset and Control Unit — Control unit contains loudspeaker, plus controls for adjustment of "volume" and "squelch." Also, "transmit," "receive," and "stand-by," signal lights.

NEW RCA FM TRANSMITTERS AND RECEIVERS

for Emergency Communications

ABOVE: FM Transmitter and Receiver for mobile or low-powered station use — Ultra-modern circuits, construction, and styling. Small, compact, rugged.

WRITE TODAY FOR 24-PAGE BOOKLET DESCRIBING THIS EQUIPMENT

Please send me complete information about your new FM Transmitters and Receivers for emergency communications. Address: Emergency Communications Equipment Section, Radio Corporation of America, Camden, New Jersey.

Name..

Title..

Employed by..

Street Address..

City and State..

RCA

RADIO CORPORATION OF AMERICA

RCA VICTOR DIVISION • CAMDEN, N. J.

in Canada, RCA VICTOR COMPANY LIMITED, Montreal

6AT6
RCA
6AU6
RCA
6BA6
RCA
6BE6
RCA
12AT6
RCA
12BA6
RCA
12BE6
RCA
35W4
RCA
50B5
RCA

9 NEW RCA MINIATURES!

INCLUDES NEW COMPACT TYPES FOR AM, FM, AND TELEVISION RECEIVERS

• Here it is—that latest addition to the peacetime line of miniatures . . . 5 new miniature types that provide performance equivalents for the popular prewar kit 12SA7, 12SQ7, 12SK7 (or 12SG7), 35Z5GT/G, and 50L6GT—and 4 other tubes introduced as performance equivalents to the 6SA7, 6SG7, 6SQ7, and 6SH7.

These 9 new RCA miniatures bring to 35 the total number of tubes in the RCA miniature line—*and all but 2 were developed by RCA engineers.*

Such RCA pioneering means two important things to you:

1. That RCA knows your needs—keeps an eye on the industry—is ever striving to give you the tubes you want when you want them.
2. As the originator of miniatures and as the largest producer of them ever since their introduction, RCA can assure you of superior tube quality and uniformity at prices that are right.

For data on these 9 new tubes, send coupon.

If you are designing radio equipment and need tube-application assistance, don't forget that the RCA application-engineering staff is always at your service for consultation. A telephone call to our nearest office, or a letter stating your problem, will do the trick. Address: Radio Corporation of America, Tube Division, Commercial Engineering Department, Section 62-41P, Harrison, N. J.

In Metals, Miniatures, or Glass Types . . .

THE FOUNTAINHEAD OF MODERN TUBE DEVELOPMENT IS RCA

AND HERE THEY ARE!

Miniatures New Types	*DESCRIPTION*	*Performance Equivalent*
★ 6AT6	Duplex-Diode High-Mu Triode	6SQ7
★ 6AU6	RF Amplifier Pentode with Sharp Cutoff	6SH7
★ 6BA6	RF Amplifier Pentode with Remote Cutoff	6SG7
★ 6BE6	Pentagrid Converter	6SA7
★ 12AT6	Duplex-Diode High-Mu Triode	12SQ7
★ 12BA6	RF Amplifier Pentode with Remote Cutoff	12SG7 (or 12SK7)
★ 12BE6	Pentagrid Converter	12SA7
★ 35W4	Half-wave High-Vacuum Rectifier	35Z5GT/G
★ 50B5	Beam-Power Amplifier	50L6GT

MAIL THIS TODAY FOR FREE DATA SHEETS

RCA, Commercial Engineering Department,
Section 62-41P, Harrison, N. J.

I'd like all the data available on the 9 new RCA miniatures announced in your December advertisement. Please rush me a complete set of data sheets, including ratings, curves, drawings, etc.

Name ....

Position ....

Company ....

Address ....

City .... Zone .... State ....

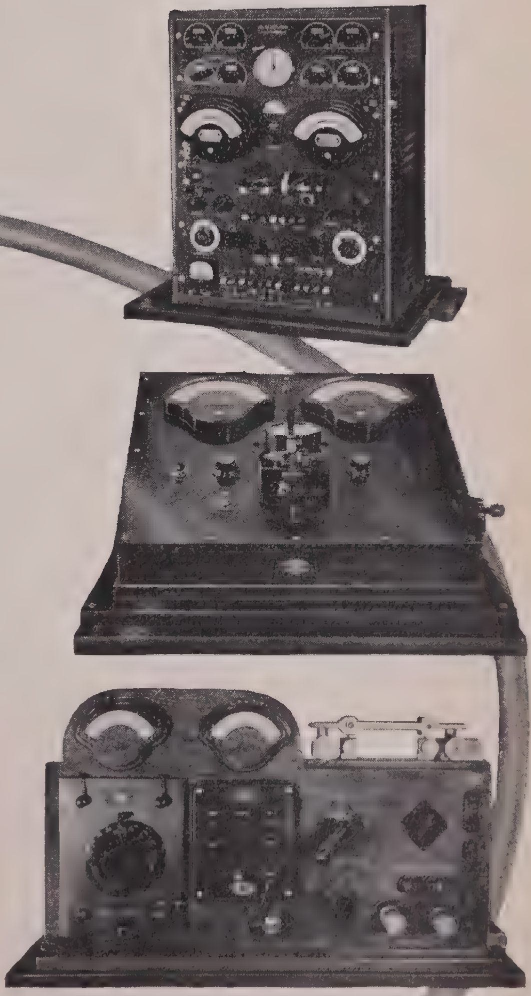
RCA

62-6136-41

RADIO CORPORATION OF AMERICA

TUBE DIVISION · HARRISON, N. J.

Start RIGHT AND YOU End RIGHT!

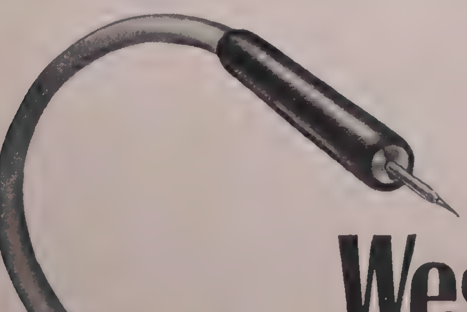


with WESTONS ON ALL PRODUCTION TEST-STANDS AND INSPECTION EQUIPMENT!

With the race for markets in full swing, electrical manufacturers are eliminating costly production bottlenecks by providing *uniform dependability* in testing procedure all along the line. From the inspection of purchased components right through to final product inspection, they insure accurate testing by using instruments they can trust.

And it's easy to insure *measurement dependability* at every step, because there are WESTONS for every testing need . . . including types for all special test-stand requirements, as well as a broad line of multi-range, multi-purpose test instruments. These compact, multi-purpose testers often afford new simplicity and economies in testing procedure, while assuring the dependability for which WESTONS are renowned.

Literature describing the complete line of WESTON panel and test instruments is freely offered . . . Weston Electrical Instrument Corporation, 578 Frelinghuysen Avenue, Newark 5, New Jersey.



Weston *Instruments*

ALBANY • ATLANTA • BOSTON • BUFFALO • CHICAGO • CINCINNATI • CLEVELAND • DALLAS • DENVER • DETROIT • JACKSONVILLE • KNOXVILLE • LOS ANGELES • MERIDEN
MINNEAPOLIS • NEWARK • NEW ORLEANS • NEW YORK • PHILADELPHIA • PHOENIX • PITTSBURGH • ROCHESTER • SAN FRANCISCO • SEATTLE • ST. LOUIS • SYRACUSE
In Canada, Northern Electric Co., Ltd., Powerlite Devices, Ltd.

Draftsman Wanted

Also

Designer
Detailer
Tracer and
Engineer

●

We are one of the largest manufacturers of a wide variety of communication and electronic equipment in the world, fully prepared and ready to go ahead with a very ambitious, expansion program as quickly as we are permitted. There will be unlimited possibilities for creative, ambitious men to advance to key positions both in research development and production field.

At present, we are producing vital equipment for our fighting forces.

●

Good Starting Salaries
Exceptionally fine working conditions.

Apply Personnel Office
8 A.M. to 5 P.M.

FEDERAL TELEPHONE & RADIO CORP.
the Mfg. unit of the International Tel. & Tel. Corp.
591 Broad St., Newark, N.J.
WMC Rules Observed

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
330 West 42nd Street, New York 18, N.Y.

RADIO ENGINEERS

For design and development of home radio receivers. Can use men with experience in design laboratories on communications and entertainment receiving sets. Call or write Mr. Frank A. Hinners, Vice President in Charge of Engineering, Air King Products, 1523 63rd St., Brooklyn 19, N.Y.

SENIOR ENGINEER

Expanding New York radio and television manufacturer, requires man with several years thorough experience in design and production of home radio receivers. Not a temporary but a permanent opportunity. High pay. Excellent future. U. S. Television Manufacturing Corp., 106 Seventh Ave., New York 11, N.Y.

ENGINEERS

A large midwest manufacturer has immediate openings for the following engineers:

Senior Development Engineers. Openings in domestic radio and television receiver development for two Senior Radio Project Engineers and one Mechanical Engineer.

Assistant Chief Engineer. To assume complete supervision of household and auto radio receiver development. Extensive prewar experience in above lines imperative. Television receivers will be in our line.

Research Engineers. Two research men with prewar development experience in the radio receiver or television field.

All inquiries will be treated as confidential. Reply to Box 406.

ENGINEERS

With experience in audio circuit work for engineering development on hearing aids and associated devices. Experience on hearing aids not actually necessary. Apply to Chief Engineer, Zenith Radio Corporation, 6001 West Dickens Avenue, Chicago 39, Illinois, giving details of education, age and experience. Excellent opportunities for interested personnel.

ENGINEER OR TECHNICIAN

To write articles on radio (particularly FM) and television, including maintenance. Fee basis. Collaboration possible for beginners. Lieutenant Myron Eddy, 295 Broadway, New York 7, N.Y.

(Continued on page 52A)

ENGINEERS

For Design Work
on Radio Receivers,
Audio Amplifiers,
Television

●

Men with substantial commercial experience wanted, preferably those having Degrees in Electrical or Communications Engineering. Write, giving details of experience and salary expected, to:

FREED RADIO CORPORATION

Makers of the Famous Freed-Eisemann Radio-Phonograph

200 Hudson Street
New York 13, N.Y.

IF YOU ARE A

SENIOR ELECTRONIC ENGINEER

Who Can Design
For Production
YOU WILL LIKE
THIS POSITION

If you have a thorough working knowledge of tubes and circuits in frequencies to 100 kc. and would like a permanent position with an alert Chicago firm, this may be your opportunity. We will pay from $5000 to $6000 a year for a man under forty who can design low frequency equipment, and who wants responsibility. There is plenty of room and opportunity for advancement. Please tell us all about yourself in your first letter—state your age, experience, education, etc. Enclose a small snapshot of yourself.

BOX 407

The Institute of Radio Engineers
330 West 42nd Street
New York 18, N.Y.

TRANSPORTATION
PETROLEUM
CONSERVATION
PROTECTION
MOTION PICTURES
UTILITIES
NEWSPAPERS

The New Bendix-Pacific

VHF MOBILE RADIO COMMUNICATION SYSTEMS

Bendix—famous for many recent new developments in radio for industry and home—now announces advanced VHF radio communication systems for mobile use.

The new systems, which offer greatly improved performance characteristics and economy of operation, are available for use in three of the Very High Frequency bands—30-44 megacycles, 72-76 megacycles and 152-162 megacycles.

All the equipment has been designed to use low input power. Any extended service can be provided with relay facilities.

For five years Pacific Division's radio research group has been working with military experts on advanced designs of VHF communication and control systems, and the new equipment embodies the latest developments for commercial application. In addition to new performance characteristics, the Bendix-Pacific equipment has been designed for compactness, long life and for easier service and maintenance.

Whatever application of radio voice communication you are planning, Pacific Division is interested in explaining the advantages of the new equipment to you. Pacific Division, Bendix Aviation Corporation, 11676 Sherman Way, North Hollywood, California.

Sales Engineering offices in New York and St. Louis.

One of the new VHF communication units combining transmitter and receiver in a single cabinet for either mobile or fixed station installation.

Pacific Division
Bendix Aviation Corporation
VHF COMMUNICATION SYSTEMS

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★ **CREI *Offers***

A Planned Program

of

GROUP

TRAINING

To Modernize the Technical Knowledge of Your Employees Through Modern Training for Professional Self-Improvement

If your organization is engaged in any phase of radio-electronics, your technical personnel must know FM—television — Ultra High Frequency Techniques, and all other phases of war-developed electronics technology; and, of course, a thorough and complete knowledge of the fundamentals of practical radio-electronics engineering.

The CREI "Employers' Plan" for group training will:

1. *Increase the technical abilities of your radio-electronics personnel.*
2. *Enable them to perform their duties more efficiently and in less time.*
3. *Increase the value of their services to your organization.*

No company time is required for this training . . . it is accomplished by spare-time, home study, and is tailored to meet each individual organizations requirements.

The CREI "Employers' Plan" is useful for the up-grading of technical personnel in manufacturing, AM, FM, and television broadcasting, communications, industrial electronics, including the following:

Engineers
Engineering Aides
Laboratory Assistants
Inspectors
Testers
Technicians
Field Servicemen
Installers
Maintenance Men

Your request will promptly bring an outline of the plan, as now in use with other organizations, and intimate details will follow when your particular needs are known. No obligation or cost, of course.

CAPITOL RADIO
Engineering Institute
E. H. RIETZKE, President

Home Study Courses in Practical Radio-Electronics Engineering for Professional Self-Improvement

Dept. PR-10, 3224—16th St., N.W.
WASHINGTON 10, D.C.

Contractors to the U. S. Navy—U. S. Coast Guard — Canadian Broadcasting Corp. Producers of Well-trained Technical Radiomen for Industry.

POSITIONS OPEN

(Continued from page 50A)

RADIO ENGINEER

Several years experience on radio receivers and associated equipment desirable. Circuit development and mechanical design. Knowledge of production test procedure and test equipment. Opportunity in new division in established firm. Premier Crystal Laboratories, Inc., 63 Park Row, New York 7, N.Y.

ACOUSTICAL ENGINEER

Preferably with E.E. degree to work in the field of microphones, phones, and supersonics. Our employees know of this advertisement. Box 399.

SEVERAL ENGINEERING POSITIONS

Progressive midwestern capacitor manufacturer has permanent key positions for top notch executive calibre men as follows:

Electrolytic Development Engineer

Capable of designing and supervising installation of equipment. To take complete charge of laboratory and supervise production quality control. Must have previous experience with etching and formation processes.

Electrical Engineer

Should have broad background of theory and practice in small electrical parts or equipment manufacturing. Position at present that of coordinating engineering problems of field sales with laboratory, engineering and manufacturing departments. Will have wide latitude of authority and report directly to management. To the right man, position will lead to that of Chief Electrical Engineer. Experience in capacitor field is advisable.

Electrical Engineer Power Factor Improvement

This key position for a new department requires an electrical engineer with specific experience in power factor improvement problems. Technical writing ability is important. The right man probably would have gained his experience with a public utility or manufacturer of heavy power equipment. He must be qualified to create and supervise an entire department for sales of capacitors used in power factor improvement. He will be given assistance of a competent staff of capacitor engineers but will be required to design and arrange for manufacture of associated power factor equipment. Sales experience will be helpful but not essential.

Applicants are requested to outline experience, education, present and previous earnings and salary requirements. All replies will be held in strictest confidence. Our own engineers know of this advertisement. Address Box 401, Proceedings of the I.R.E., 330 West 42nd Street, New York 18, New York.

(Continued on page 54A)

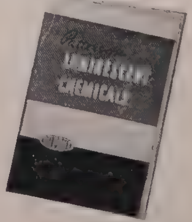
Quality Phosphors

FOR A VARIETY OF NEEDS

Patterson high-grade phosphors, produced singly or in blends, meet a variety of specific needs. For more than thirty years, the Patterson Screen Division of the Du Pont Company has manufactured superior fluorescent materials. Principal component of world-famed Patterson Screens for precision x-ray work, these phosphors facilitate important radiographic diagnosis.

Manufactured under exacting conditions, Patterson phosphors are distinguished for uniformity of emission, color and grain size. Shades, grain structures and life of afterglow are assured.

Throughout the war period, Patterson Luminescent Chemicals served makers of cathode ray tubes for television and radar equipment. Now they are available for similar applications and research work. A descriptive booklet sent upon request. Patterson Screen Division of E. I. du Pont de Nemours & Co. (Inc.), Towanda, Pa.



Patterson Luminescent Chemicals

DU PONT

BETTER THINGS FOR BETTER LIVING
...THROUGH CHEMISTRY

250 Watts of Broadcast Satisfaction

Collins 12Z Remote Amplifier

A high quality four channel remote amplifier, a.c.-d.c. powered. The d.c. source consists of self-contained batteries which take the load automatically in case of a.c. line failure. Gain, approximately 95 db. Frequency response, 30-12,000 c.p.s. ± 1 db. Power output, 50 milliwatts. Weight, with batteries and carrying case, 32 pounds.

The new Collins 300G-1 AM broadcast transmitter is an operator's ideal. Its components are the finest available, with very high safety factors, *and all are completely and immediately accessible.* Replacements, if necessary, are just a quick, simple one-man job!

Circuit design, physical arrangement, and workmanship throughout, meet the superior standards which station engineers have come to expect of Collins engineering.

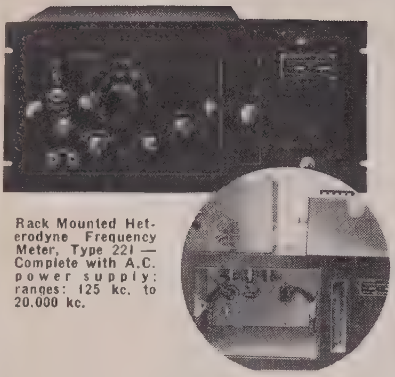
The nominal power output of the 300G-1, 250 watts, can be reduced to 100 watts by means of a switch on the control panel. The response is flat within ± 1.0 db from 30 to 10,000 cycles. Distortion is less than 3% up to 100% modulation.

Tell us about your plans. We will be glad to study them with you and make recommendations covering requirements for your entire station, AM or FM, and of any power. Collins Radio Company, Cedar Rapids, Iowa; 11 West 42nd Street, New York 18, N. Y. In Canada, Collins equipment is sold by Collins-Fisher Limited, Montreal.

FOR BROADCAST QUALITY, IT'S . . . COLLINS

CARDWELL

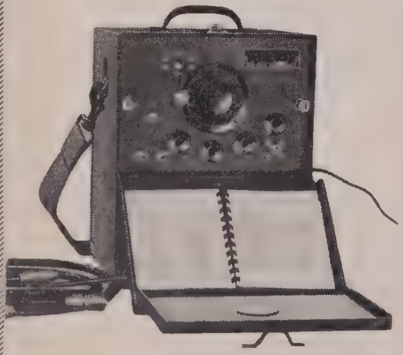
skill and facilities responsible for this "SPECIAL ORDER" Equipment are available for your design and production problems . . .



Rack Mounted Heterodyne Frequency Meter, Type 221 — Complete with A.C. power supply; ranges: 125 kc. to 20,000 kc.

Similar Frequency Meter in Portable Case—Battery operated (or A.C. power supply can be substituted); ranges of various Frequency Meters:

Type 221	125 kc. to 20 Mc.
Type 173	90 Mc. to 450 Mc.
Type 174	20 Mc. to 250 Mc.
Type 175	85 Mc. to 1000 Mc.



These are instruments of accuracy and precision as manufactured for government service by the makers of Cardwell variable and fixed air capacitors, which have long been a "Standard of Comparison."

The Cardwell engineering and manufacturing personnel and facilities, so skillfully employed on this equipment, are available for any production or design problem in the field of electronics.

Address Inquiries to:
DEPT. M, CONTRACT MANUFACTURING DIVISION

THE ALLEN D. CARDWELL MANUFACTURING CORP.
Main Office: 81 Prospect St., Brooklyn 1, N.Y.
Factories: Plainville, Conn. – Brooklyn, N.Y.

POSITIONS OPEN

(Continued from page 52A)

TELEVISION ENGINEERS

Design experience in the development of television cameras, terminal equipment or transmitters. State experience and salary desired. Apply in person or in writing to Personnel Department, Raytheon Manufacturing Company, Inc., Communications Division, 60 E. 42nd St., New York 17, N.Y.

SALES ENGINEER

For sale of new and interesting electronic device. Must have had previous sales experience in radio or other electronic equipment. A real opportunity for right man with large international organization. Write qualifications in detail to Box 400.

DIRECTOR OF ENGINEERING

Smaller firm with national organization and established post-war business in electronic, audio and electro-acoustic fields has opening for engineer in charge of development and design. Salary liberal. Must have engineering degree and practical experience. Design of audio and electro-acoustic systems. Replies confidential. State education, experience record, patents, etc. Photograph if available. Box 398.

SALES ENGINEER

Manufacturer of materials for radio components needs sales engineer with good technical background to appraise and work on technical details in the sales of their products. Headquarters New York. Box 402, Proceedings of the I.R.E., 330 West 42nd Street, New York 18, New York.

ENGINEERS

Opportunities are offered by an expanding, progressive engineering organization to first-class Development, Communication and Radio Engineers with extensive prewar experience. Write: Maguire Industries, Inc., Electronics Division, Personnel Dept., 342 W. Putnam Ave., Greenwich, Conn.

RADIO PARTS OF "TAYLOR LAMINATED PLASTICS"

WHETHER OR NOT your

Post-war radio has the

Tone, selectivity and quality

Appearance to give it a head start

On competition depends to a large

Extent upon how closely the

ENGINEERS who design it follow

Modern developments, especially

In the field of LAMINATED PLASTICS

Where great strides have been made

During the WAR YEARS in the

Development of new and better

Materials and methods.

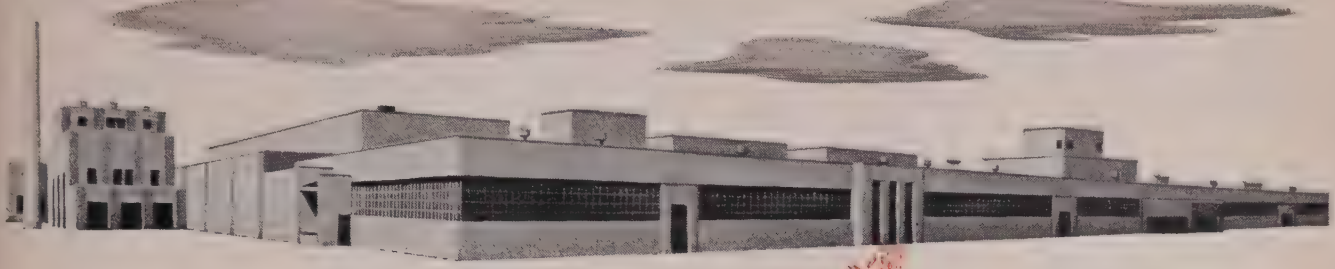
Like to KNOW MORE? Write us today.

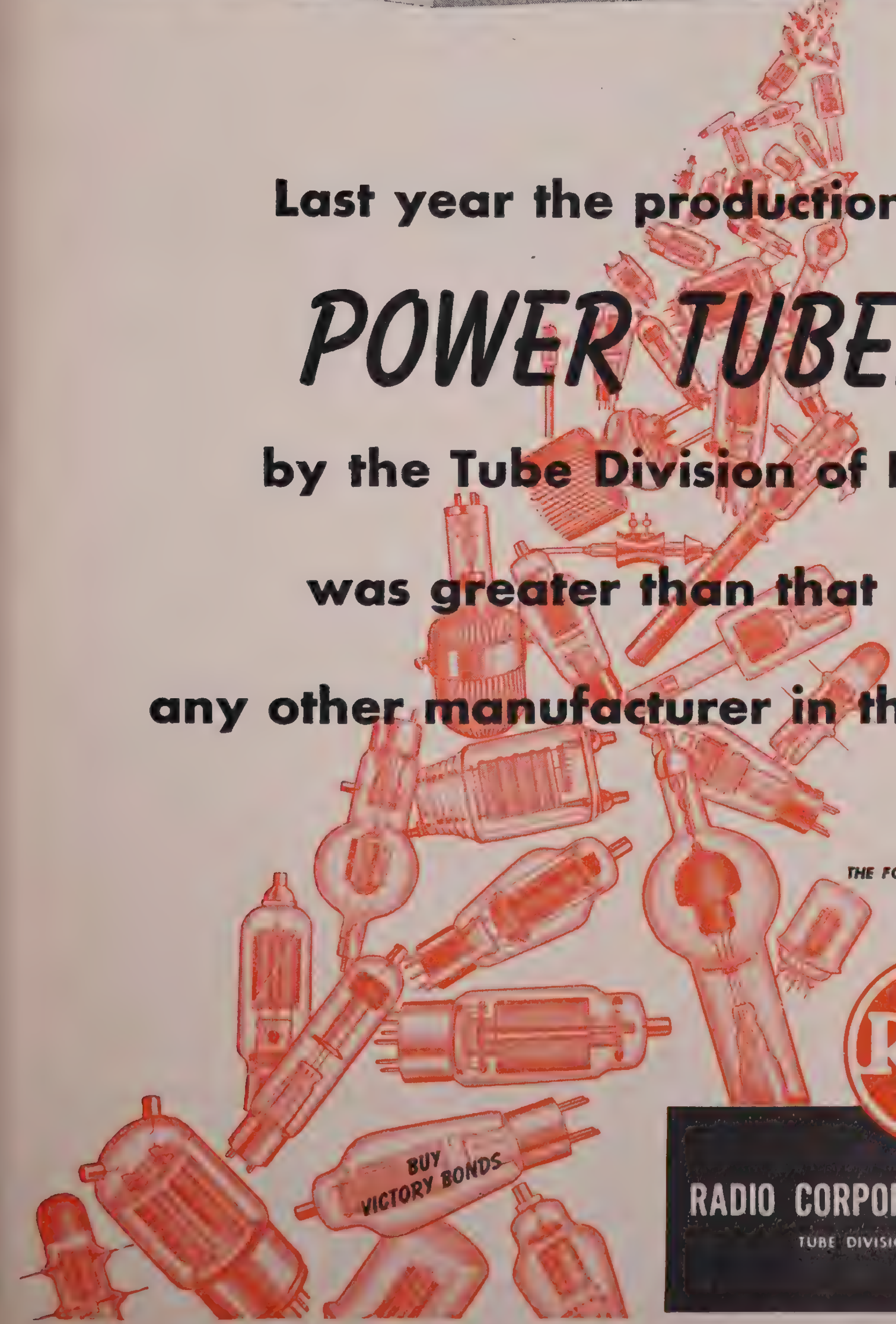
This two-part radio insulator is typical of Taylor's ability to mass-produce accurate parts. Each is stamped in a single operation, one from Vulcanized Fibre and the other from a grade of Phenol Fibre noted for its high dielectric strength. Whatever combination of qualities you may need, it's a good bet that Taylor can give it to you.

TAYLOR FIBRE COMPANY

LAMINATED PLASTICS: PHENOL FIBRE · VULCANIZED FIBRE
Sheets, Rods, Tubes, and Fabricated Parts
NORRISTOWN, PENNSYLVANIA
OFFICES IN PRINCIPAL CITIES
Pacific Coast Headquarters:
544 S. SAN PEDRO STREET, LOS ANGELES

RCA Tube Plant, Lancaster, Penna.
RCA manufactures cathode-ray tubes, as well
as power tubes, in this modern plant.





Last year the production of

POWER TUBES

by the Tube Division of RCA

was greater than that of

any other manufacturer in the world

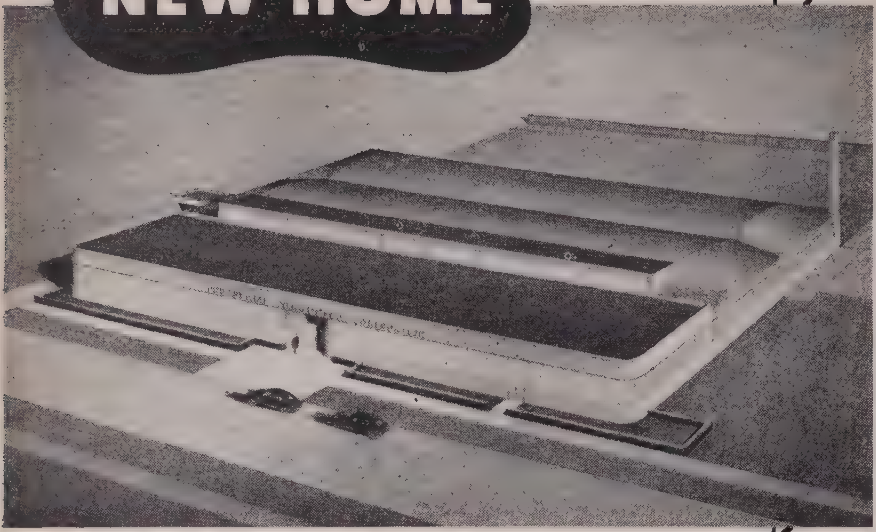
*THE FOUNTAINHEAD OF MODERN
TUBE DEVELOPMENT IS RCA*

RCA

62-6136-37

RADIO CORPORATION OF AMERICA
TUBE DIVISION • HARRISON, N. J.

Industrial Condenser Corporation's
NEW HOME



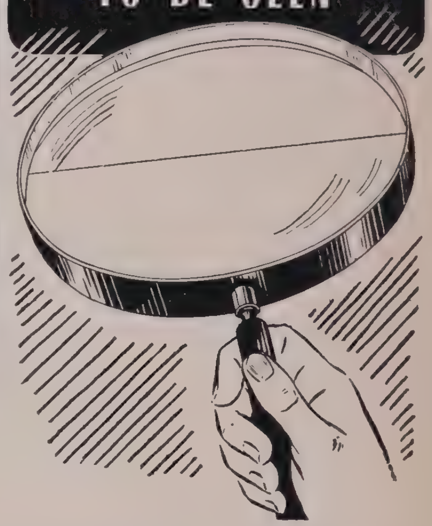
THE WORLD'S MOST
MODERN CONDENSER PLANT
with these outstanding features

★ 1,000,000 VOLT RESEARCH LABORATORY
★ VERY LATEST PRODUCTION EQUIPMENT
★ SPECIALIZED WAR-LEARNED TECHNIQUES

From this NEW ultra-modern factory come capacitors carefully engineered and accurately produced. Staffed by skilled engineers and backed by 16 years of technical progress, Industrial Condenser Corp. is supplying capacitors for every application. If your specifications call for Electrolytic, Paper, Oil, or Motor capacitors, look to Industrial Condenser Corporation.

INDUSTRIAL CONDENSER
CORPORATION
3243-65 NORTH CALIFORNIA AVENUE, CHICAGO 18, U. S. A.
District Offices in Principal Cities
PAPER, OIL AND ELECTROLYTIC MOTOR CAPACITORS

WIRE SO FINE
IT MUST BE
Magnified
TO BE SEEN



WOLLASTON PROCESS
Wire as small as

$\frac{1}{100,000}$ { OF AN INCH IN DIAMETER

. . . available in Platinum and some other Metals

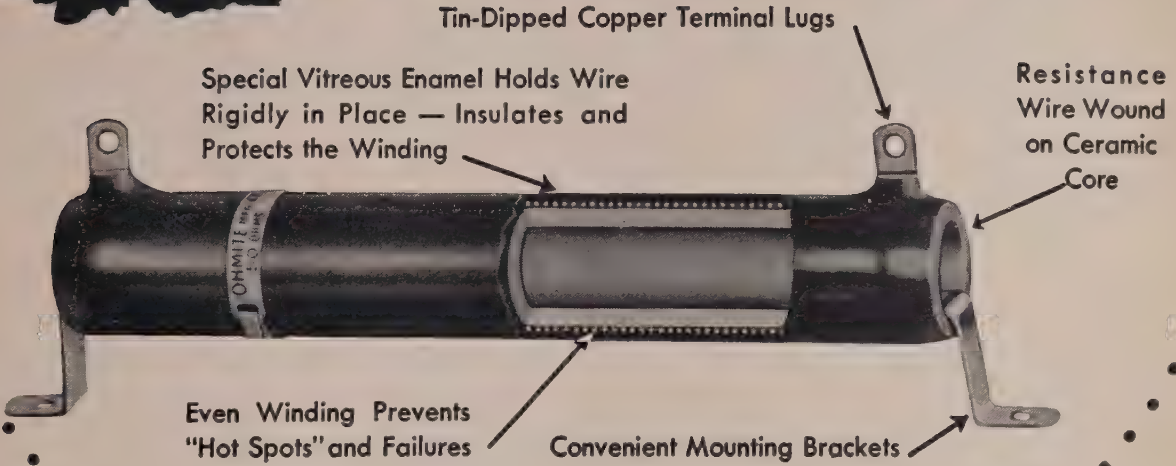
.00001″ is less than 1/30 the diameter of the smallest wire die commercially available. Yet our Wollaston Process wire (drawn in a silver jacket) closely meets your specifications for diameter, resistance and other characteristics.

●

This organization specializes in wire and ribbon of smaller than commercial sizes and closer than commercial tolerances. *Write for List of Products.*

SIGMUND COHN & CO.
SINCE 1901
44 GOLD STREET NEW YORK 7

Why OHMITE Resistors PERFORM SO DEPENDABLY


Tin-Dipped Copper Terminal Lugs
Special Vitreous Enamel Holds Wire Rigidly in Place — Insulates and Protects the Winding
Resistance Wire Wound on Ceramic Core
OHMITE
Even Winding Prevents "Hot Spots" and Failures
Convenient Mounting Brackets

Here you see a few of the important features that insure long life and trouble-free service in every Ohmite Resistor. In the lug type illustrated above, the resistance wire is both mechanically locked and brazed to copper terminal lugs to assure perfect electrical connection. Time-proved Ohmite vitreous enamel construction dissipates heat rapidly . . . withstands humidity. Today, Ohmite Resistors are extensively used in the Armed Forces, Industry, Communications, Research. *Made in a wide range of types and sizes* in stock and special units for every need. Consult Ohmite engineers on your resistor problem.

OHMITE MANUFACTURING COMPANY
4861 FLOURNOY ST., CHICAGO 44, U. S. A.

Write on company letterhead for Industrial Catalog and Engineering Manual No. 40. Gives helpful information on resistors, rheostats, chokes, tap switches.

Be Right with OHMITE
RHEOSTATS • RESISTORS • TAP SWITCHES

ALNICO-5
"MIRACLE METAL"
for
SPEAKERS

Cinaudagraph Speakers are known the world over for tone, stamina, engineering perfection and design. Consider the use of Alinco 5, newest miracle metal that gives 4 times the performance without weight or size increase, add this to scores of other Cinaudagraph Speaker achievements and you have the reasons why Cinaudagraph Speakers are "The Finest Speakers in all the World."

Cinaudagraph Speakers
A DIVISION OF Aireon
3911 SOUTH MICHIGAN AVENUE, CHICAGO
"The Finest Speaker in all the World"

3223
ELECTRONIC Engineering COMPANY

Ready... Willing... Able...

Ready to take on any transformer production job . . . our complete new manufacturing facilities are second to none in the industry.

Willing to handle any transformer design or development job in elaborate new laboratories staffed with engineers of unusual experience.

Able to ship, quickly, "one million or one" of any transformer you order. Write, wire or phone us for fast efficient service.

SPECIALIZED *Transformer* ENGINEERS

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of the
Proceedings of the I.R.E.
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Each issue is priced at $1.00 but members of I.R.E. are given a 25% discount and accredited public and college libraries a 50% discount. The following issues are offered subject to prior sale.

1913-1915 Volumes 1-3
Quarterly

1913	Vol. 1	January (a reprint)

1916-1926 Volumes 4-14
Bimonthly

1917	Vol. 5	April, June, August, October
1918	Vol. 6	None
1919	Vol. 7	December
1920	Vol. 8	April, June, August, October, December
1921	Vol. 9	All 6 issues
1922	Vol. 10	All 6 issues
1923	Vol. 11	February, April, June, August, December
1924	Vol. 12	August, October, December
1925	Vol. 13	April, June, August, October, December
1926	Vol. 14	All 6 issues

1927-1938 Volumes 15-26
Monthly

1927	Vol. 15	April, May, June, October, December
1928	Vol. 16	February to December, inc.
1929	Vol. 17	April, May, June, November
1930	Vol. 18	January, April to December, inc.
1931	Vol. 19	All 12 issues
1932	Vol. 20	January, March to December, inc.
1933	Vol. 21	All 12 issues
1934	Vol. 22	All 12 issues
1935	Vol. 23	All 12 issues
1936	Vol. 24	January to April, inc., June
1937	Vol. 25	February to December, inc.
1938	Vol. 26	All 12 issues

1939-1945 Volumes 27-33
Monthly

New Format—Large Size

1939	Vol. 27	All 12 issues
1940	Vol. 28	All 12 issues
1941	Vol. 29	All 12 issues
1942	Vol. 30	January, February, April, October to Dec., inc.
1943	Vol. 31	February to December, inc.
1944	Vol. 32	January, February, July to December, inc.
1945	Vol. 33	All issues

Remittance should accompany your order.

Please do not send cash except by registered mail

THE INSTITUTE OF RADIO ENGINEERS, INC.
330 West 42nd Street,
New York 18, N.Y.

CLAIM STAKING

Hallicrafters and Very High Frequency

Based on the facts in the case, Hallicrafters can stake out a very strong claim to leadership in the very high frequency field. The facts include such things as the Model S-37, FM-AM receiver for very high frequency work. The Model S-37 operates from 130 to 210 Mc.—the highest frequency range of any general coverage commercial type receiver.

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Further developments in this direction can soon be revealed—adding further support to Hallicrafters claim to continued supremacy in the high frequency field.

HALLICRAFTERS NEW $600,000 HOME NOW UNDER CONSTRUCTION.

hallicrafters RADIO

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AND ELECTRONIC EQUIPMENT • CHICAGO 16, U. S. A.

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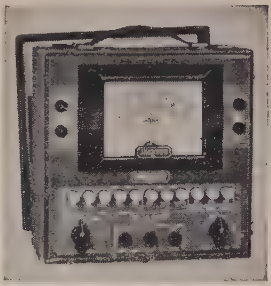
Start with any 1 Jackson Instrument to Build a Balanced Testing TEAM!

Condenser Tester
Model 650A—Measures Capacity, Power Factor and Leakage



Sensitive Multimeter
Model 642—20,000 ohms per volt—complete ranges

Electronic Multimeter
Model 645—A new Jackson instrument of advanced design

Tube Tester
Model 634—Uses exclusive Jackson "Dynamic" Test Method

Multimeter
Model 643—1000 ohms per volt. Push key range selection

Test Oscillator
Model 640—Accurate to ½%, covers full frequency range

IT'S A PLUS VALUE of the Jackson line. Each instrument is engineered and manufactured for long accurate life, as today's users know—but every one is carefully matched in appearance, dimensions and finish as well.

Start with whichever Jackson instrument you need first. Add to it as occasion demands. Your foresight will be repaid with a matched and balanced set of instruments built to give you testing results that you just can't get with hit-or-miss assemblies. Plan now to equip your shop with these Jackson instruments. See your distributor.

Service Lab
Assembly of *Standard Size* Jackson Instruments

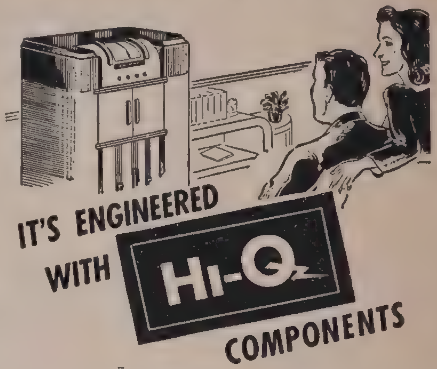
BUY WAR BONDS AND STAMPS TODAY

JACKSON

Fine Electrical Testing Instruments

JACKSON ELECTRICAL INSTRUMENT COMPANY, DAYTON, OHIO

DEPENDABLE PERFORMANCE

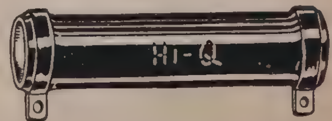


Dependable performance in any electronic equipment is the sum of little things. And when those "little things" are Hi-Q components you can rest assured that performance will be thoroughly dependable. Hi-Q components are available *in any desired quantities* to your specifications. Send for samples and complete data.

CERAMIC CAPACITORS

CI Type: axial leads
CN Type: parallel leads

Made of titanium dioxide (for temperature compensating types). Tested for physical dimensions, temperature coefficients, power factor and dielectric strength.



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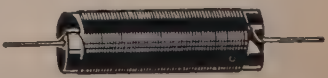
Fixed Type

Immediately available in standard ratings or precision built to any tolerance or value.



CHOKE COILS

Sturdy construction. Insulated or uninsulated. Quantity production available at once.

ELECTRICAL REACTANCE CORPORATION
FRANKLINVILLE, N. Y.

How SKILSAW Protects Equipment "*in-the-Package*"

Moisture "in-the-package" causes vast damage to precision equipment.

Skilsaw, Inc. prevents this damage by means of Jay Cee Silica Gel, the ideal drying agent. A few small bags of the Jay Cee Gel are included within tightly sealed packages of electric handsaws, drills, and other tools. The gel has amazing power to absorb and hold moisture. Thus the air in the package is kept thoroughly dry and damage from rust or corrosion is avoided.

This practice is being followed by more and more manufacturers of metal products. The cost of the gel is a mere trifle in comparison with the possible damage from rust or corrosion.

JAY CEE
GEL
A SUPERIOR DEHYDRANT
PROTECTING SHIPMENTS
FROM THE ELEMENTS

Jay Cee Silica Gel has wide application in the Air Conditioning, Refrigeration, and Chemical industries. It is clear white; passes a rigid section test; meets exacting Government specifications; is strictly a quality product.

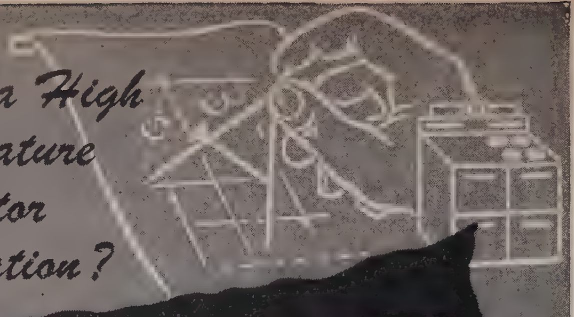
A few excellent Jay Cee Silica Gel sales territories are still open to jobbers. Write for details.

LIET CHEMICALS, LTD., INDUSTRY AVENUE, JOLIET, ILLINOIS

AY CEE SILICA GEL

A superior dehydrant

Have you a High Temperature Capacitor Application?



PLASTICONS Operate at 256° F.

Plasticon Types AS are silicone impregnated plastic film dielectric capacitors. They will operate from —80° F. to +256° F. with less than 10% capacitance change. Plasticons are small in size as they are not "voltage derated."

CP

Condenser Products Company

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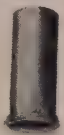
1. This deep-drawn shell, 1 13/16"d x 4 3/8" is drawn without expensive annealing, by the Goat Precise-Formed Process.

2. Stamped thread ring shown above is induction brazed to the shell.



3. A thread is cut on the ring, completing the product.

You, too, can save by

GOAT PRECISE-FORMED METAL STAMPINGS

Induction brazing and machining operations can save you money by replacing many types of machine parts, castings and assemblies. New techniques in deep drawing eliminate the necessity of in-between annealing. As a result, we can produce stampings, as well as assemblies of stampings, that would have been regarded as impossible prior to the war.

GOAT METAL STAMPINGS

Samples or Prints analyzed without obligation

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1

GOAT METAL STAMPINGS, INC.
Affiliate of The Fred Goat Co., Inc.
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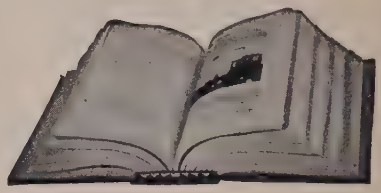
Binders

for the Proceedings

Protect your file of copies against damage and loss



Binders are available for those who desire to protect their copies of the PROCEEDINGS with stiff covers. Each binder will accommodate the twelve monthly issues published during the year. These binders are of blue Spanish grain fabricoid with gold lettering and will serve either as temporary transfers or as permanent binders. They are so constructed that each individual copy of the PROCEEDINGS will lie flat when the pages are turned. Copies can be removed from the binder in a few seconds and are not damaged by their insertion.

Available for both the old, small size PROCEEDINGS or the new, large size (1939 to date).

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either size (specify which)
Postpaid to all countries

You may have a volume number or your name stamped in gold for 50 cents additional.

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RADIO ENGINEERS, INC.
330 West 42nd Street,
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Build the competitive advantages of longer life, better performance into your products . . . with

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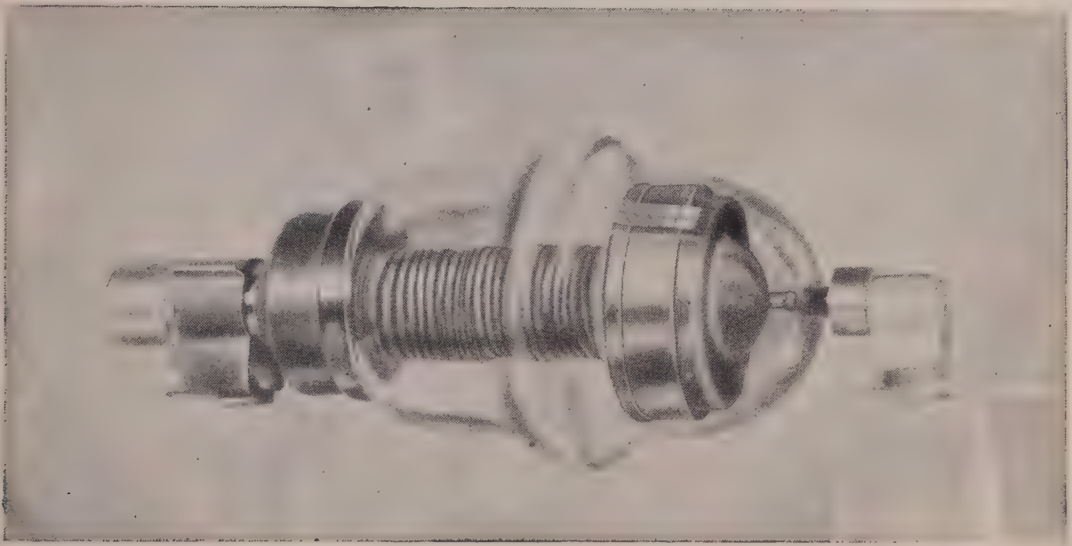

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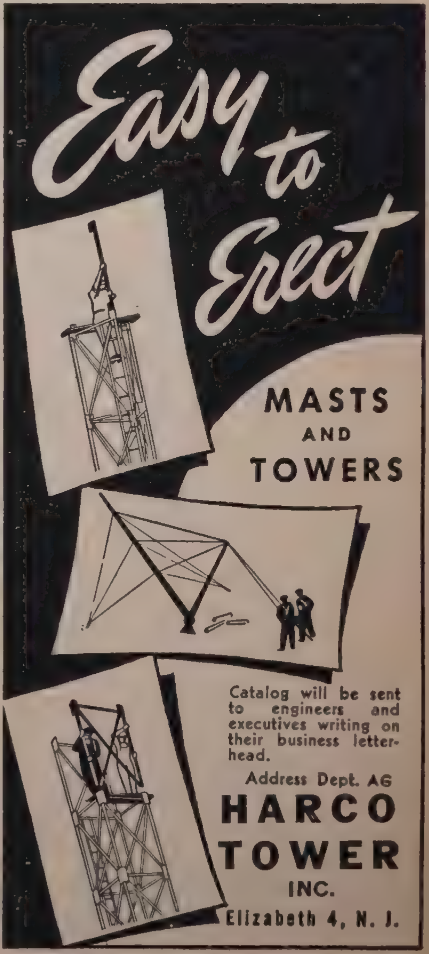

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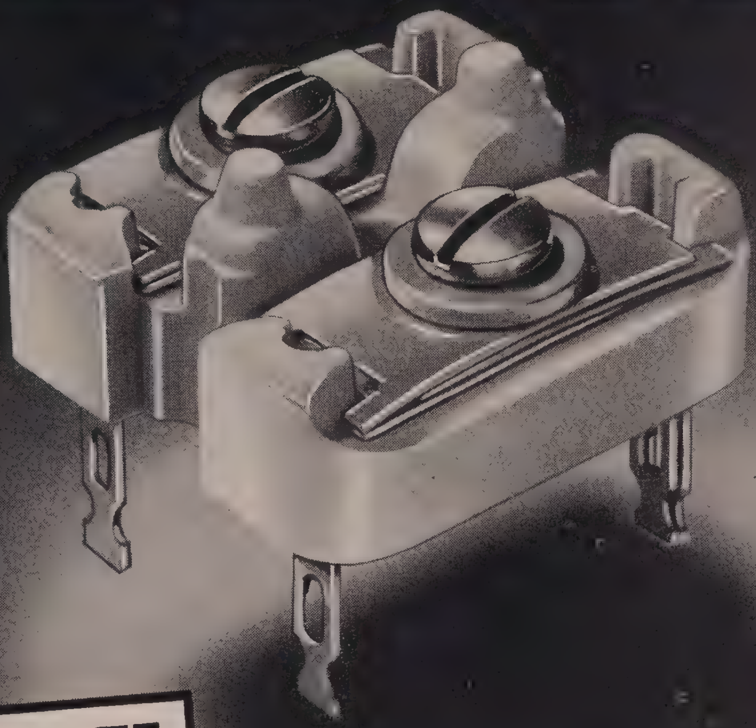

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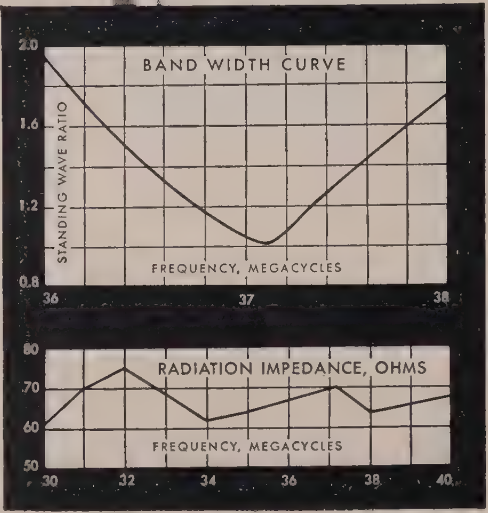

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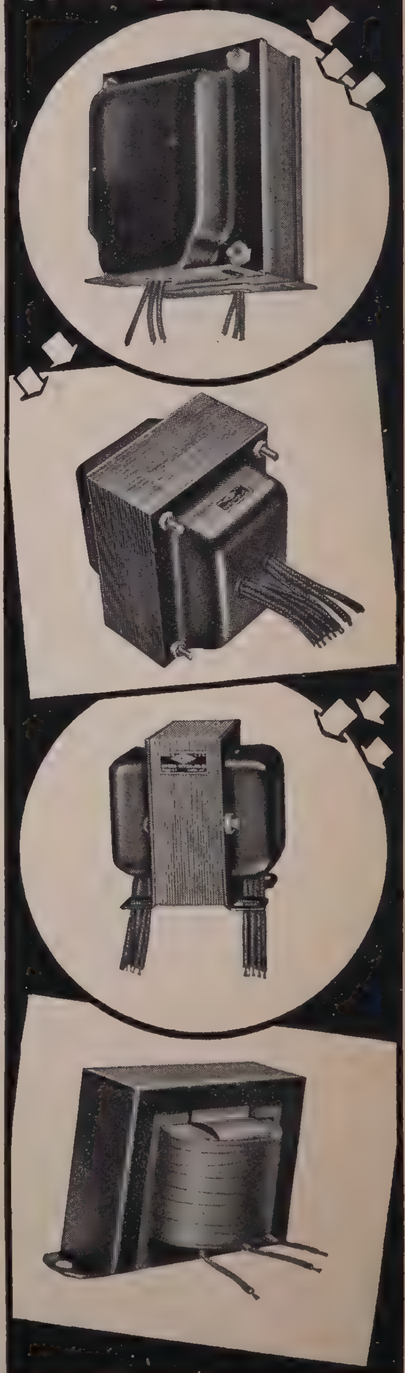

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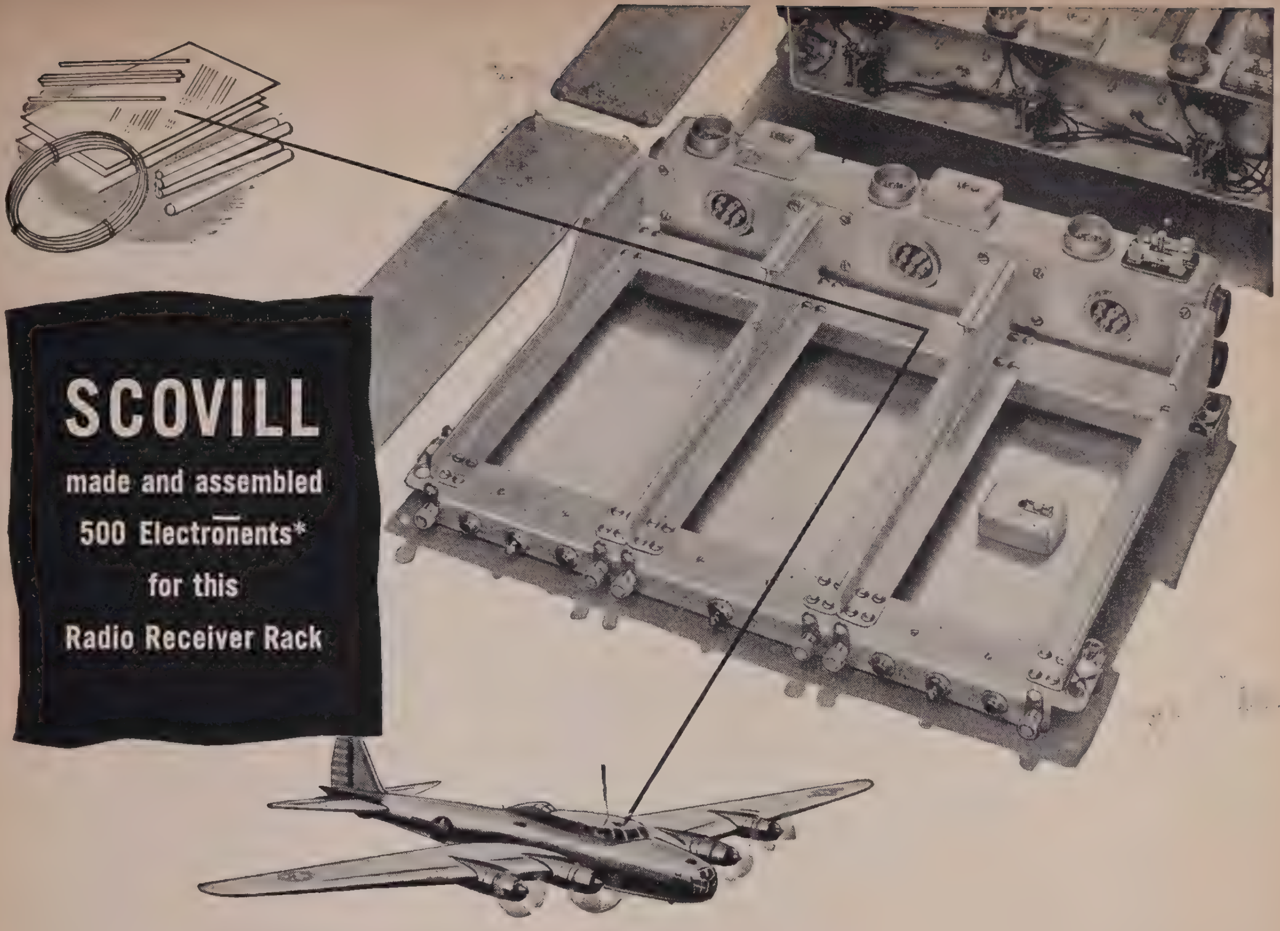

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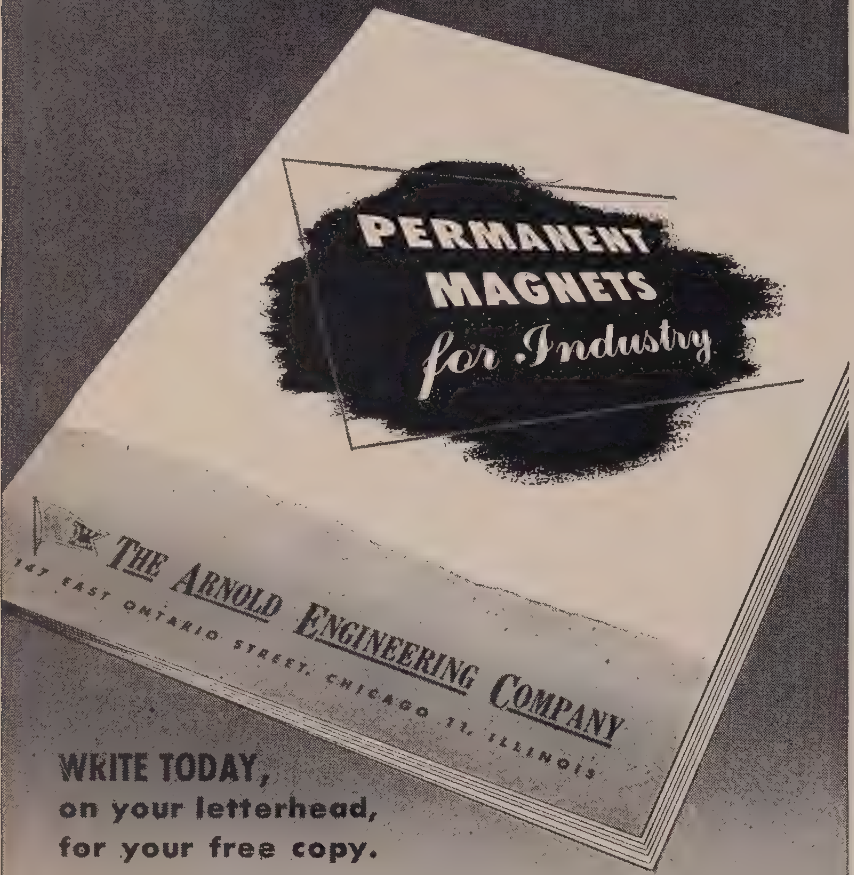
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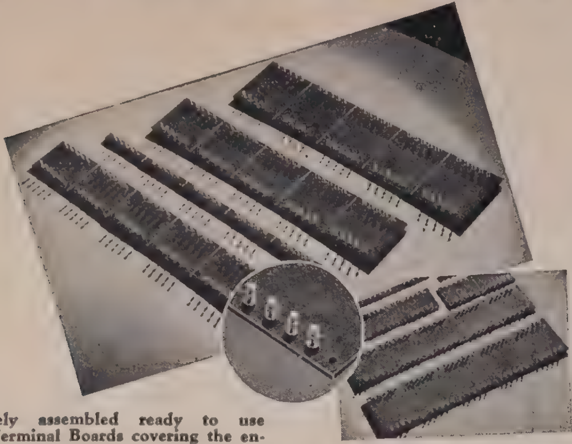

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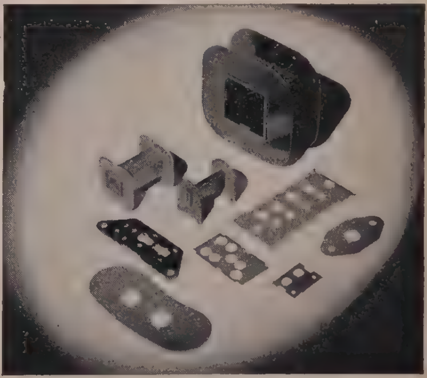

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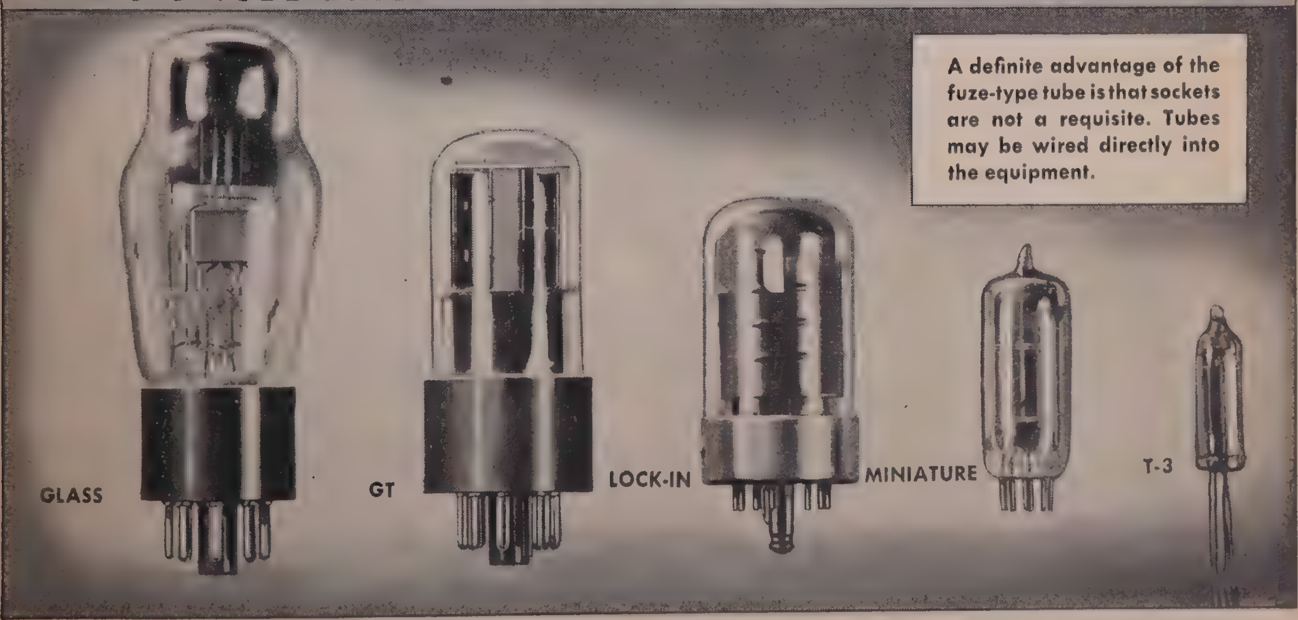
THE T-3 TUBE LINES UP WITH OTHER FAMOUS SYLVANIA TUBES



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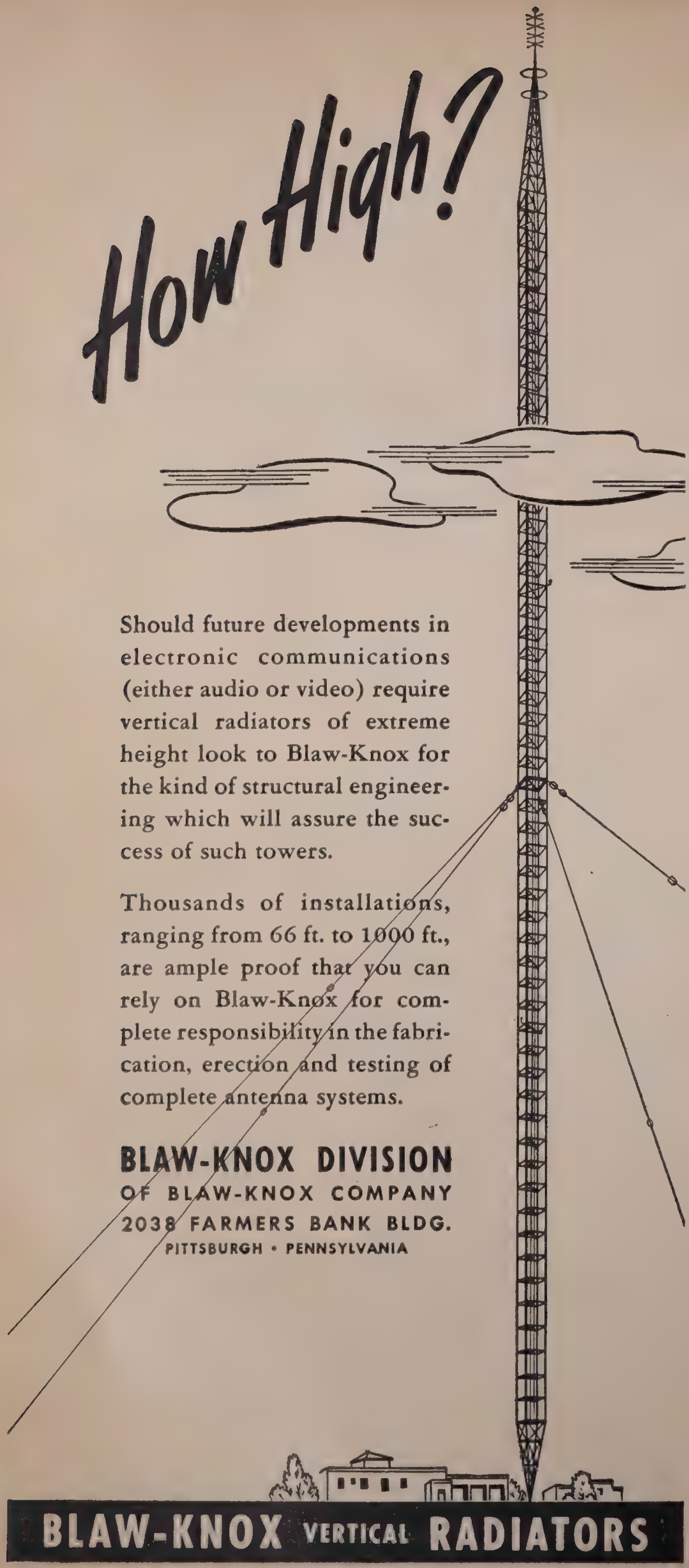

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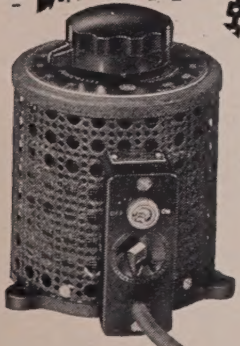
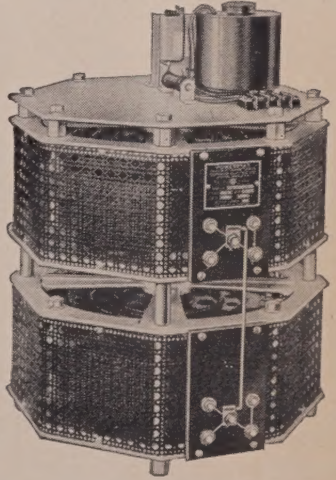


The majority of the standard POWERSTAT variable transformers are carried in stock for immediate shipment. For instance; if a smooth continuously adjustable output voltage is required from a 115 volt line, POWERSTAT type 116 is available with an output range of 0 to 135 volts. Its current rating is 7.5 amperes.



For control of POWERSTATS from push-button stations or automatic controllers, motor-driven POWERSTATS are offered in the same capacities as manually-operated units. Only a few days is necessary to make delivery of any standard model.

SECO'S increased production of SECO Automatic Voltage Regulators means that the unequalled performance of the SECO Regulator in maintaining a constant output voltage regardless of variations in input voltage or output load current is available for all requirements in the range of 1 to 100 KVA.

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American Transformer Co. 57A
Amperex Electronic Corp. 17A
Andrew Company 76A
Arnold Engineering Co. 78A
Audio Development Co. 38A
Automatic Manufacturing Corporation 75A

Ballantine Laboratories, Inc. 74A
Alfred W. Barber Laboratories 74A
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Brown Engineering Company 72A

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Eicor, Inc. 42A
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Eitel-McCullough, Inc. 88A
Electrical Reactance Corporation 68A
Electronic Engineering Co. 66A
Electronics Research Publishing Co. .. 84A

Electrons, Inc. 36A
Erie Resistor Corp. 25A

Federal Tel. & Radio Corp. ..15A, 37A, 50A
Freed Radio Corporation 50A

General Electric Co. 8A
General Radio Company ....Cover IV
Goat Metal Stampings, Inc. 70A
General Instrument Corp. 11A

Hammarlund Mfg. Co., Inc. 12A
Hallicrafters Co. 47A, 67A
Harco Tower Inc. 74A
Heintz & Kaufman, Ltd. 6A
Hewlett-Packard Co. 26A
Hytron Radio & Electronics Corporation 29A

Industrial Condenser Corp. 64A
International Resistance Company 23A
International Tel. & Tel. Corp. ....
.........................15A, 37A, 50A
Instruments Publishing Co. 84A

Jackson Electrical Instrument Co. ... 68A
Jennings Radio Mfg. Co. 73A
Joliet Chemicals, Ltd. 69A

Kaar Engineering Company 19A

Lewis Electronics 32A

Machlett Laboratories, Inc. 59A
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Measurements Corp. 79A
Carl H. Miller Engr. Co. 86A
Mycalex Corporation of America 41A

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National Co., Inc. 39A
National Union Radio Corp. 27A
National Vulcanized Fibre Co. 80A
Newark Electric Co., Inc. 74A
New York Transformer Co. 45A

Ohmite Mfg. Co. 65A
M. F. M. Osborne 86A

Albert Preisman 86A
Press Wireless, Inc. 55A

Radio Corp. of America 32B, 32C, 63A
Rauland Corp. 31A
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Remler Co., Ltd. 80A
Bernard Rice's Sons, Inc. 43A
Richardson Company 46A

Scovill Manufacturing Company 77A
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Triplett Electrical Instrument Co. 56A
Tung-Sol Lamp Works, Inc. 10A

United Transformer Corp. ....Cover II

Victoreen Instrument Co. 78A

Western Electric Co. 2A, 3A
Westinghouse Electric Corp. 5A
Weston Electrical Instrument Corp. .. 49A

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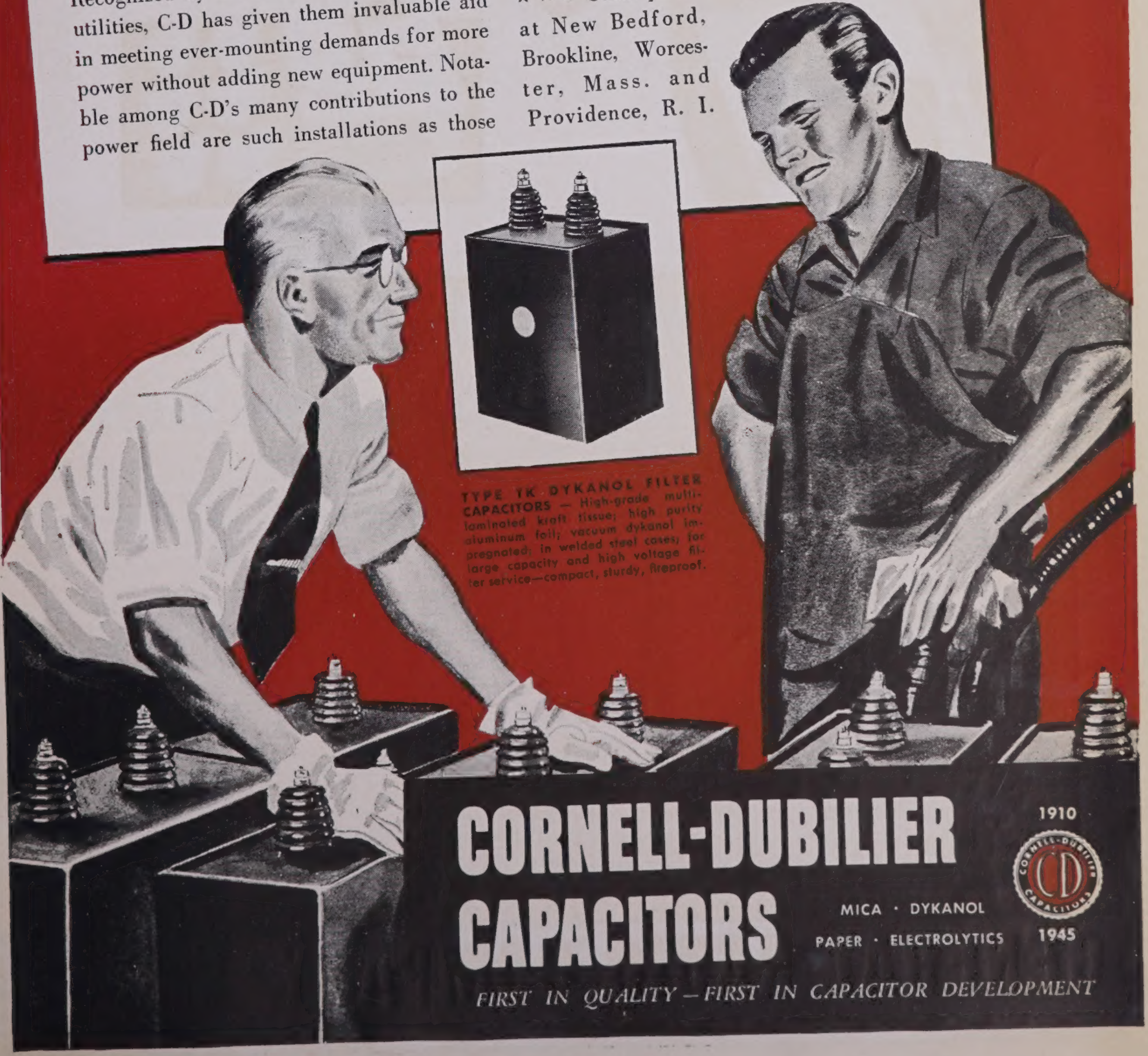

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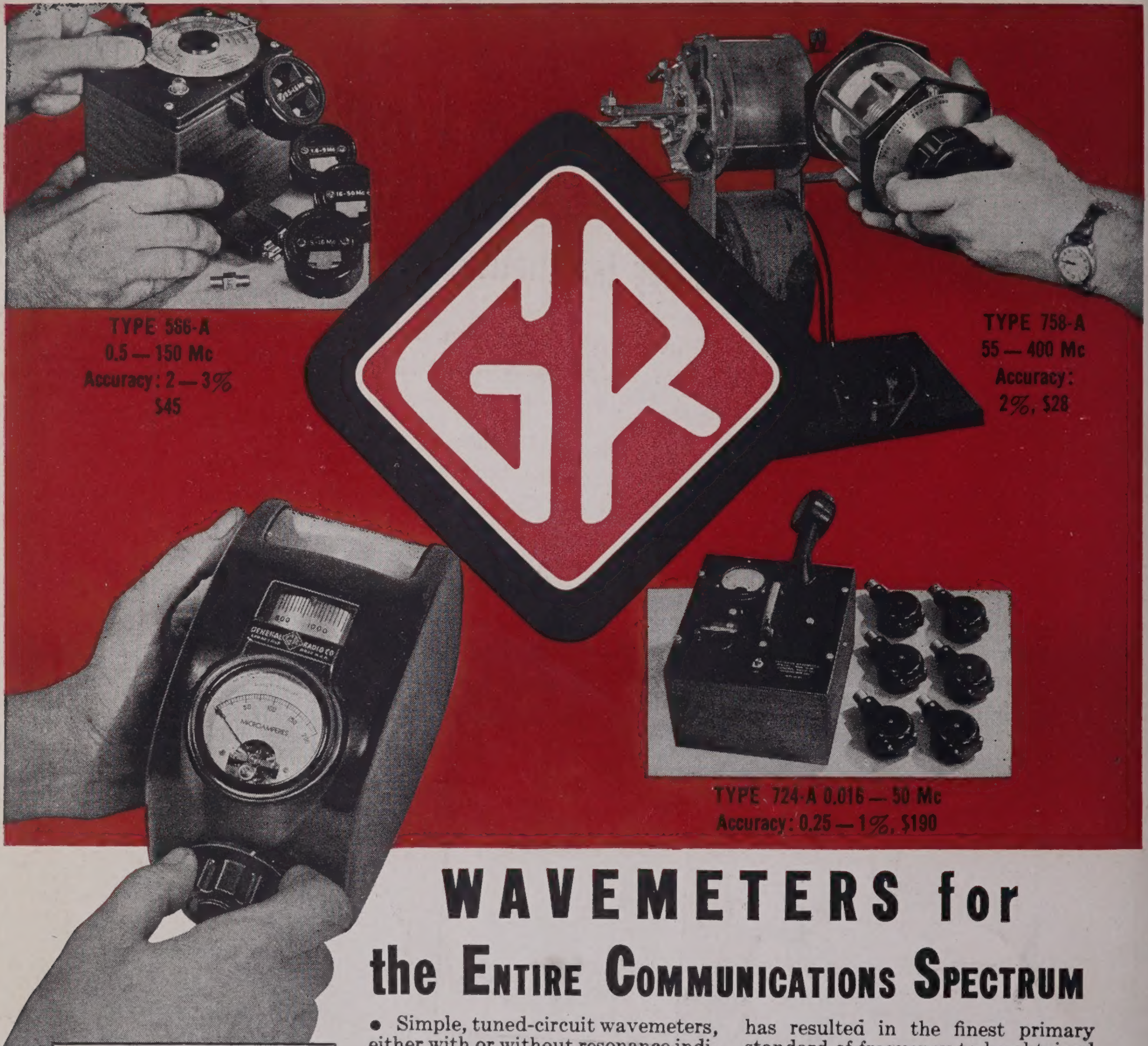

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